WIRELESS ENGINEER

THE JOURNAL OF RADIO RESEARCH & PROGRESS

APRIL 1954

VOL. 31 No. 4 THREE SHILLINGS AND SIXPENCE

Model

UNIVERSAL AVOMETER

20,000 ohms per volt *plus* AUTOMATIC Overload Protection

Produced in response to a demand for a high sensitivity version of the world-famous Universal AvoMeter, this model incorporates the traditional design features of its predecessors, so highly valued for simplicity of operation and compact portability.

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25V.	1mA.	25V.	2.5A.	0—2,000Ω (using internal batteries)
100V.	10mA.	100V.	10A.	0—200,000Ω (using internal batteries)
250V.	100mA.	250V.	—	0—20MΩ (using internal batteries)
1,000V.	1A.	1,000V.	—	
2,500V.	10A.	2,500V.	—	0—200MΩ (using external batteries)

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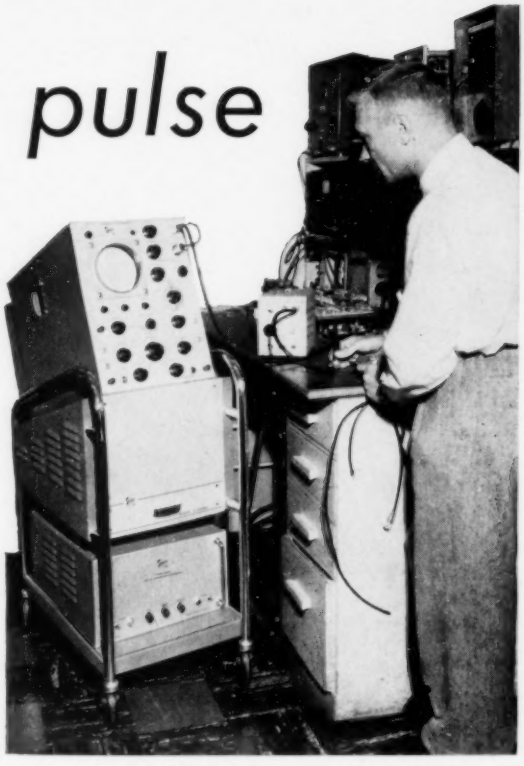
...The Tektronix TYPE 517



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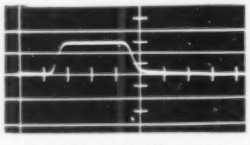
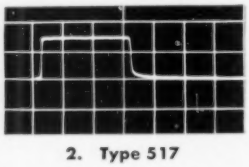
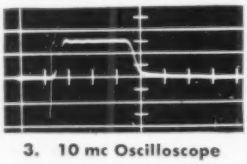


1. 10 mc Oscilloscope

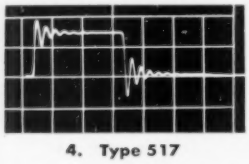


2. Type 517

0.3 μsec pulse with cable improperly terminated



3. 10 mc Oscilloscope



4. Type 517

The waveform photographs at left illustrate the capabilities of the Type 517 in one application dealing with extremely fast-rising pulses. The same 0.3 μsec pulse is shown in all four photographs—first with a three-foot length of 93-ohm coaxial cable properly terminated (photos 1 and 2), and below with improper output termination (photos 3 and 4). Notice in photo 4 how clearly the Type 517 shows the effects of improper termination on the pulse.

If your project is one that requires an oscilloscope with the faithful transient response, fast sweeps, and high accelerating potential of the Type 517, call your Tektronix Field Engineer or Representative for complete specifications—or write to:

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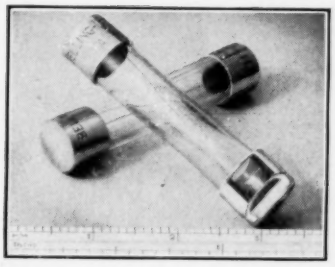
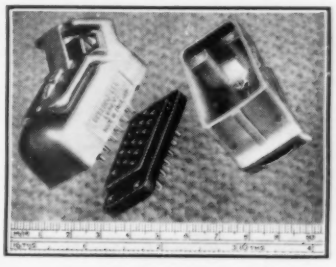
P. O. Box 831D Portland 7, Ore. Cable: TEKTRONIX

The "Belling-Lee" Page

Guide to the R.E.C.M.F. Exhibition

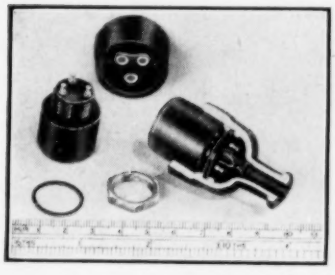




Stand No. 55

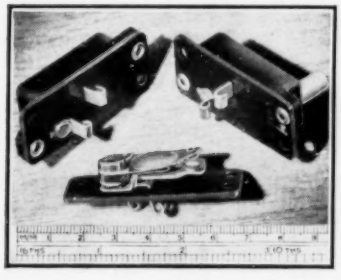
New and established lines which have been redesigned, feature prominently in the 1954 range of "Belling-Lee" components and accessories for the Electronic, Radio, and Electrical Industries.

"Unitor" and "Screenector" contacts are also available hard gold plated. This finish will be added to other lines as appropriate.



Many ratings of the well known general purpose instrument fuselink, L.1055, are now manufactured by an entirely new technique which bonds caps, glass, and filament into one unit, caps being so securely held that they will not come off unless the glass is broken.

A preview of a new range of six fuseholders for Inter-Service use can be seen, and these, in addition to the existing types of sealed and neon indicating versions, will form a very comprehensive range.



Screened plugs and sockets with 4, 6, and 12-way assemblies have been introduced. Housings are interchangeable with existing screened coaxial types.

Three and four mm resilient sockets with square faced nylon moulding are exhibited. Sockets can be mounted singly or in groups.

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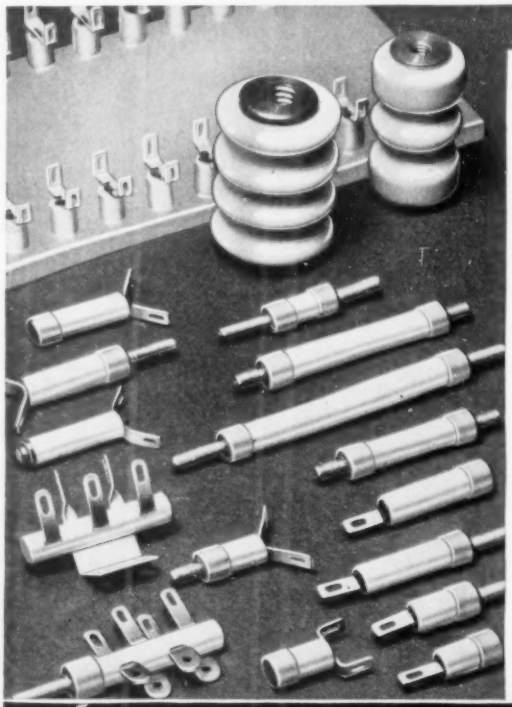

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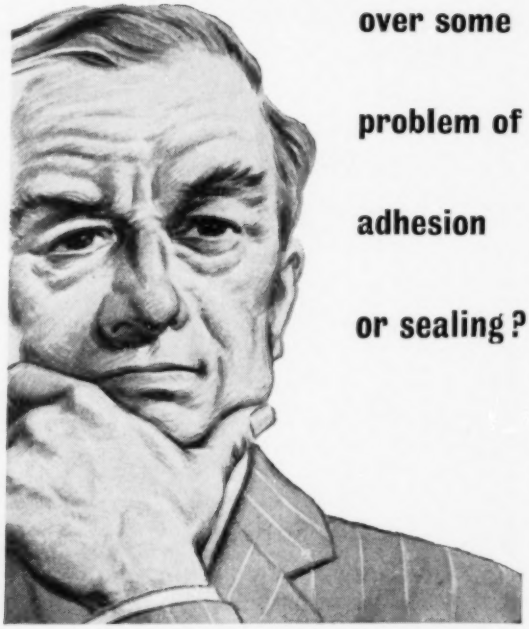

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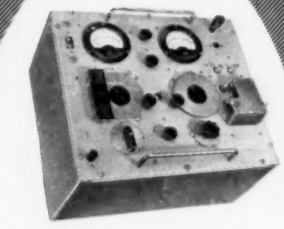
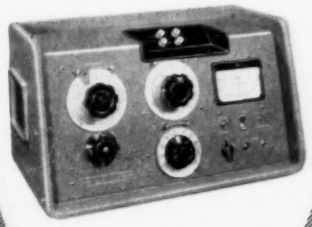


TF 329G: 50 kc/s — 50 Mc/s



TF 886A: 15 — 170 Mc/s

TF 329G; 50 kc/s—50 Mc/s; 10—500 Q; 40—450 $\mu\mu$F.
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TC 8

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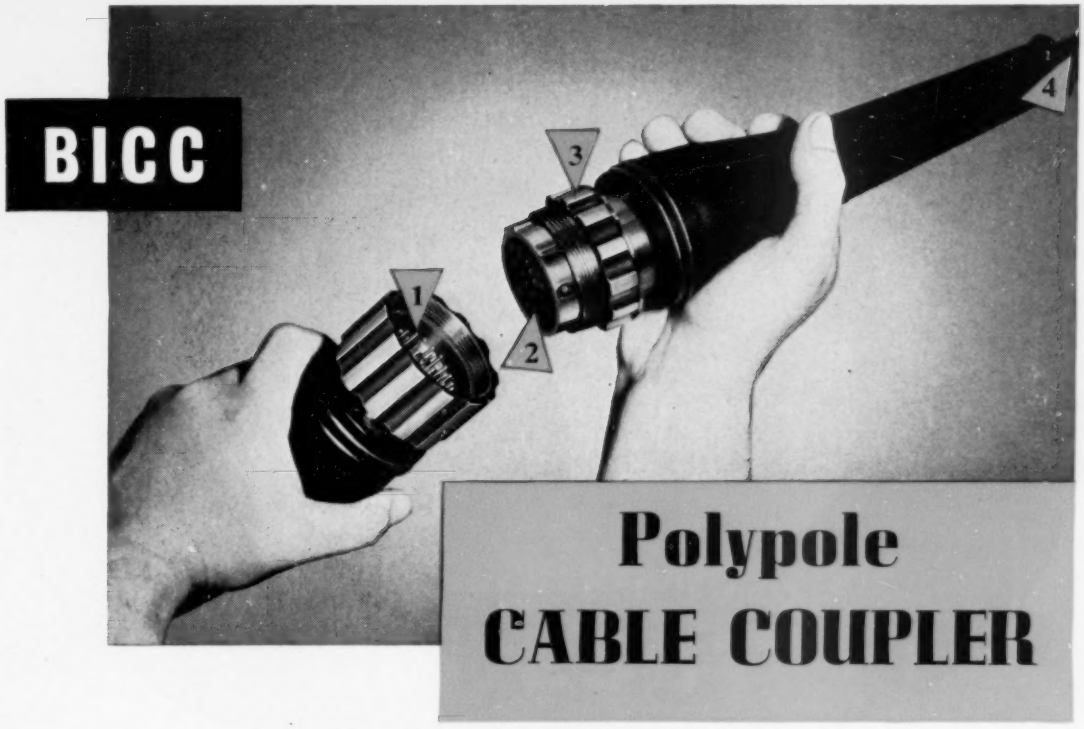

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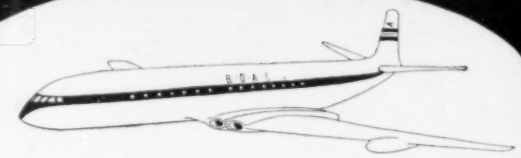

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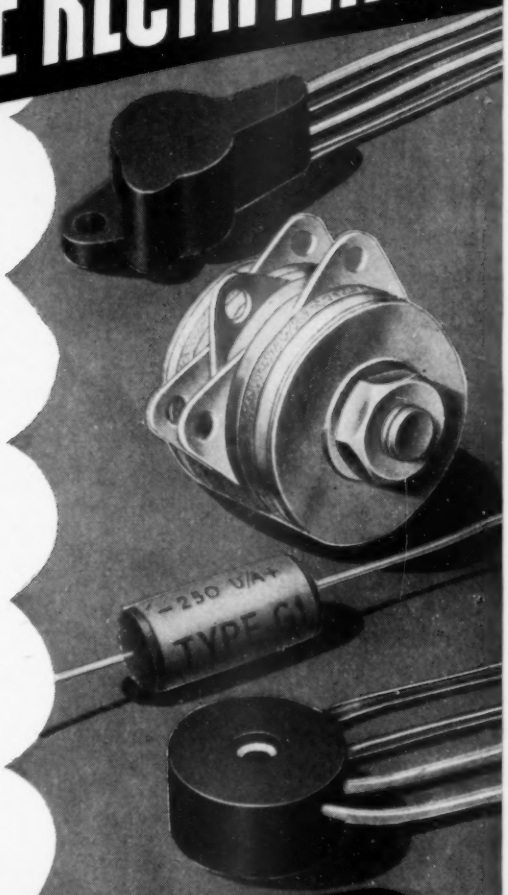

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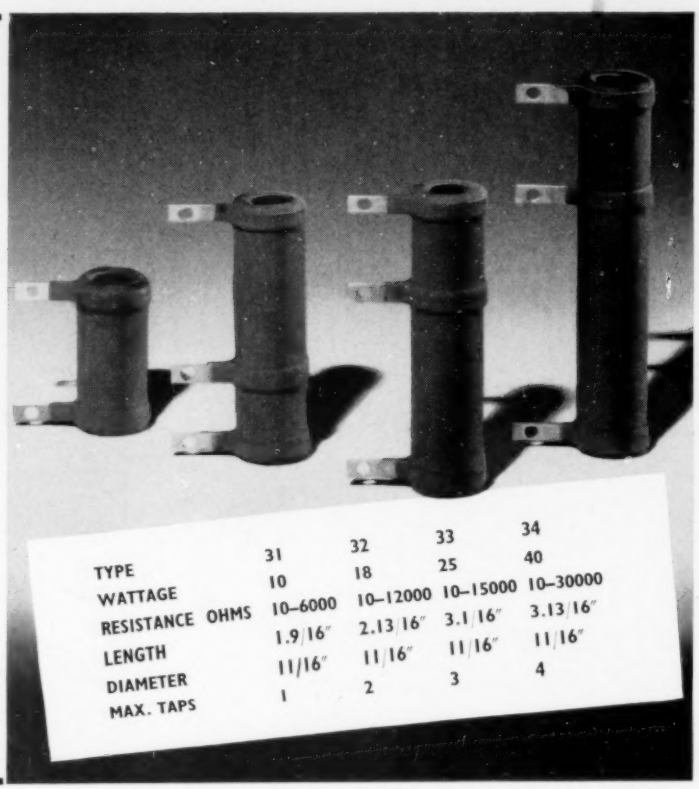

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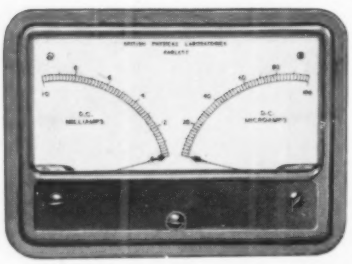

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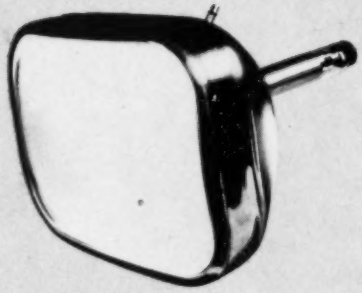

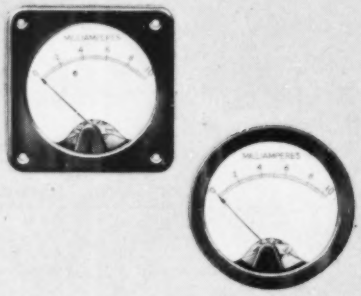

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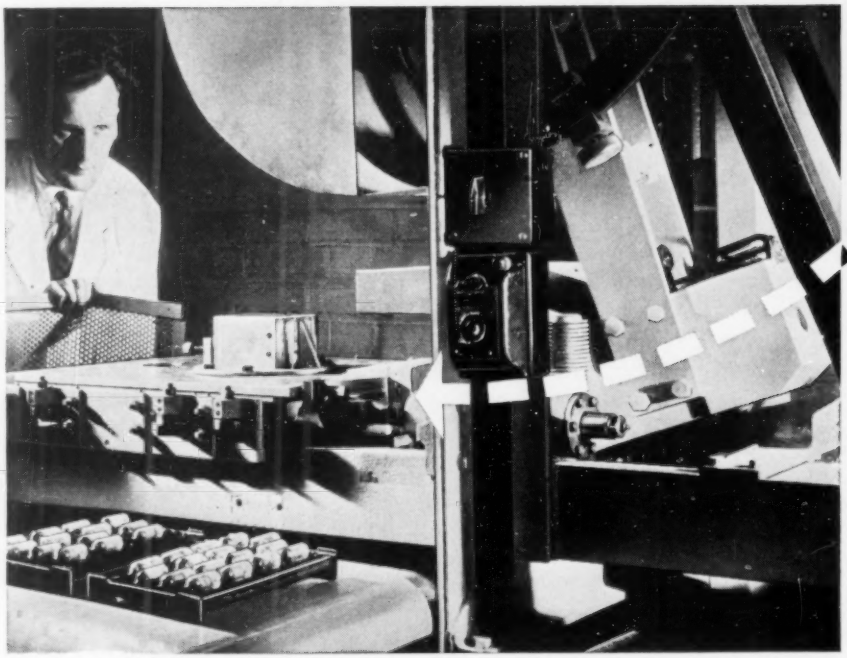

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WIRELESS ENGINEER

Vol. 31 APRIL 1954 No. 4

H. A. Lorentz

HENDRIK ANTOON LORENTZ was born on 18th July 1853, and to mark the centenary Dr. Balth van der Pol contributed a paper entitled "H. A. Lorentz and the Bearing of his Work on Electromagnetic Telecommunication" to the *Journal of the International Telecommunication Union*, which was reprinted in the November number of *Tijdschrift van het Nederlands Radiogenootschap*. Lorentz was for many years Professor at the University of Leiden; when he retired in 1923 he acted as Director of Research at the Teyler Institute, Haarlem, but continued as an Honorary Professor of Leiden until his death on 4th February 1928. At the beginning of the century he was universally regarded as the leading theoretical physicist of his time. Albert Einstein recently wrote, "I often had an opportunity to attend H. A. Lorentz's lectures which he regularly gave to a small circle of younger colleagues after his retirement from his general professorship. Everything that emanated from this supremely great mind was as clear and beautiful as a good work of art; and one had the impression that it all came out so easily and effortlessly as I have never experienced it from anyone else. If we younger ones had known H. A. Lorentz only as a great luminary, our admiration and veneration for him would already have been of an extremely special kind, but what I feel when I think of H. A. Lorentz is not covered by a long way by that veneration alone. For me personally he meant more than all the others I have met on my life's journey."

Dr. van der Pol is uniquely qualified to discuss the work of the two outstanding physicists J. J. Thomson and H. A. Lorentz, for he worked at Cambridge with the former from 1917 to 1919 and then at Leiden with the latter from 1919 to 1922. With regard to the electron he says that it was *invented* by Lorentz and *discovered* by Thomson. Lorentz obtained his doctorate in 1875 for a thesis on the application of Maxwell's equations to the reflection and refraction of light. He postulated the existence of small electrically-charged particles in 1878, and when in 1896 Zeeman, also at Leiden, showed that when a sodium flame was subjected to a strong magnetic field its spectrum lines were displaced, Lorentz deduced from Zeeman's measurements that the value of e/m of the electrons vibrating in the sodium flame was of the order of 10^7. We now know it to be $1{\cdot}602 \times 10^{-20}$ e.m. units/$9{\cdot}12 \times 10^{-21}$ gram $= 1{\cdot}75 \times 10^7$. Lorentz also predicted that the emitted light would be polarized, which Zeeman immediately confirmed. In October 1897, J. J. Thomson published his measurements of the deflection of the particles in a discharge tube by electric and magnetic fields by means of which the value of e/m could be calculated. Lorentz and Zeeman were both awarded the Nobel Prize in 1902.

Lorentz showed that in calculating the electric field at a point in a gas or mixture of gases the polarization of neighbouring molecules must be taken into account. In the ionosphere this may be negligible, but it must be allowed for when calculating the dielectric constant of the atmosphere.

In 1892 he introduced the conception of retarded potentials.

In 1905 Lorentz discussed the reciprocity theorem for linear systems and stated it in a form

somewhat more general than that originally used by Helmholtz in 1853, although, as we pointed out in the July 1943 Editorial, Helmholtz disclaimed originality and gave Green the credit for discovering the reciprocity theorem and applying it to static electricity. Helmholtz did not consider alternating currents and only mentioned the magnitude of the current, whereas Lorentz mentions amplitude and phase. His wording is as follows:—"If an e.m.f. applied at a point P in the direction h produces at a point P' a current whose component in an arbitrarily chosen direction h' has the amplitude u and phase v, an equal e.m.f. applied at the point P' in the direction h' will produce a current at P whose component in the direction h has exactly the same amplitude u and phase v." This is merely an elaboration of Helmholtz's wording to cover alternating current.

Van der Pol says that from personal conversations he knows that Lorentz had a great admiration for the work of Heaviside, and in one of his last papers he developed a generalization of a little-known theorem of Heaviside (Electrical Papers, Vol. II, p. 412). Van der Pol states the theorem as follows. Given a constant passive network at rest; if at time $t = 0$ a constant e.m.f. is suddenly applied to the network, in general, transients will occur. After a long time t_1 the transients may be considered to have died down, and only a direct current will, in general, be present in the network. The direct current (if there is any) will cause a Joulean heat dissipation at a constant rate per second. If we called W' the "pseudo heat dissipation" which would have occurred if the constant final current had been present all the time t_1, then, when the steady state has been reached, the total amount of energy A supplied by the e.m.f. exceeds the "pseudo heat dissipation" (if there is any) by twice the excess of the electric energy U over the magnetic energy T, or $(A - W') = 2(U - T)$. It follows that a condenser can only be charged from a d.c. source with an efficiency of 50%. Just as he did with Helmholtz's reciprocity theorem, Lorentz generalized this theorem of Heaviside and introduced distributed capacitances and inductances.

Lorentz made important and very fundamental contributions to the special theory of relativity. He was always deeply interested in the propagation of electromagnetic waves through a moving medium, and when in 1887 Michelson and Morley found that the velocity of light was the same whether in the direction of the earth's motion or in the opposite direction, Lorentz came to the same conclusion as Fitzgerald, viz., that matter moving at a high velocity appears to be contracted in the direction of its motion, and that this would apply to the length-measuring apparatus to a degree just sufficient to cancel the increased time that the light took to travel, and thus give an unchanged observed velocity. Lorentz also postulated that all observations made in a laboratory would be the same whether the laboratory was at rest or travelling with constant velocity through the ether. Unlike Einstein, Lorentz preferred to picture the ether as the wave medium. They both arrived at the same result and established the set of equations known as the Lorentz transformation, which are at the basis of the whole theory of relativity. Einstein went further by generalizing them and making them applicable to systems moving at a variable speed.

Van der Pol says that Lorentz was the ideal chairman for an international science congress, for he was a master of Dutch, English, French and German, and could summarize a complicated contribution in such a way that it became clear to the audience. He was not much in favour of such congresses, for he said that while they were discussing some problem, a scientist working alone in his laboratory might find the solution.

One must not confuse H. A. Lorentz with Ludvig Lorenz of Copenhagen (1829–1891), who made a number of contributions to the *Annalen der Physik*, and whose method of determining the ohm was employed in many standardizing laboratories.

G. W. O. H.

BOLTZMANN INTERFEROMETER

Interpreting Interferograms

By J. L. Farrands, B.Sc., Ph.D., A.Inst.P., and J. Brown, M.A., A.M.I.E.E.

(Imperial College of Science and Technology)

SUMMARY.—The Boltzmann interferometer is the most convenient instrument at present available for the investigation of the spectrum radiated by spark-type oscillators. It is shown that the power spectrum of the oscillator is given by the Fourier transform of the interference pattern obtained from the instrument. A typical example is examined in detail.

THERE is at present considerable interest[1,2] in extending the radio spectrum to higher frequencies; i.e., to wavelengths in the millimetre and sub-millimetre bands. Some success has been achieved in extending the frequency range of klystrons and magnetrons, the types of oscillator which have proved so successful at centimetric wavelengths,[3] but there appears to be still a use for oscillators which use sparked resonators.[4]

Such oscillators have the property that the power produced is spread over a wide range of frequencies at relatively low levels. It is essential for a full understanding of the mechanism of operation of these oscillators to determine the power spectrum; i.e., the way in which the power density varies with frequency. The instruments which can be used for such measurements are diffraction gratings, cavity resonators and interferometers. Diffraction gratings are subject to the two disadvantages of low efficiency and the difficulty of interpreting the results, because of the occurrence of second-order spectra. Apart from the constructional problems arising at short wavelengths, cavity resonators are difficult to use when the energy is spread over a wide range of wavelengths because of the occurrence of more than one resonance. Interferometers are relatively simple to use and are very efficient, the most suitable for the present purpose being that known as the Boltzmann interferometer[5,6].

The arrangement of the interferometer is shown schematically in Fig. 1. The radiation from the source is focused into a plane parallel beam by the paraboloidal mirror, A, and is directed on to the two interferometer plates, P_1 and P_2, at an angle of incidence θ. These plates are assumed to be perfect reflectors of sufficient size to give complete specular reflection of the incident radiation, which is then directed towards a second paraboloidal mirror, B. At the focus of B is placed the detector. The apparatus must be aligned in such a way that equal powers reach the detector from each of the plates. This may be done by placing P_1 and P_2 side by side so that they form a continuous surface and then adjusting the axes of A and B so that the detector output is halved when either P_1 or P_2 is removed.

MS accepted by the Editor, June 1953

When the apparatus has been adjusted as above, the interference pattern is obtained by displacing P_2 a distance s normal to the plane of its surface as in Fig. 1, and plotting the detector output, S, against s. The resulting curve is called an interferogram and from it can be deduced the required information on the power spectrum of the source. This has usually been done merely by noting the main periodicities in the interferogram but, as will now be shown, considerably more information can in fact be obtained.

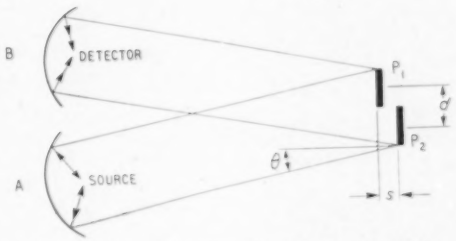

Fig. 1. *General arrangement of interferometer.*

After reflection from the plane mirrors, P_1 and P_2, the incident plane wave is effectively split into two plane waves, which travel in the same direction, but are displaced in time by an interval τ. From the geometrical theory of optics it can be shown that this time interval is given by

$$\tau = 2s/c \cos\theta \quad .. \quad .. \quad .. \quad (1)$$

where c is the phase velocity of electromagnetic radiation in free space. If the angle θ is small, say less than 10°, $\cos\theta$ is not appreciably different from unity and then

$$\tau = 2s/c \quad .. \quad .. \quad .. \quad (2)$$

Let $f(t)$ be the electric field strength at the detector caused by the reflection from P_1. Then the field strength from reflection at P_2 is $f(t-\tau)$ and so the total field strength acting on the detector is

$$F(t) = f(t) + f(t-\tau) \; .. \quad .. \quad (3)$$

If the source is a spark-type oscillator, the radiation will be in the form of a succession of damped wave trains. It is convenient in the remainder of the analysis to assume that $F(t)$ refers to only one of these wave trains. Most forms of detector give an indication proportional to the total power received; i.e.,

$$S = K \int_{-\infty}^{\infty} |F(t)|^2 dt \quad .. \quad .. \quad (4)$$

Because of the damped nature of the wave train represented by $F(t)$, it is permissible to use the limits $-\infty$ and $+\infty$ for the time integration; K is a constant of proportionality whose precise value is not required. In stating equation (4), it has been assumed that the detector responds equally to all frequencies; this restriction will be removed below.

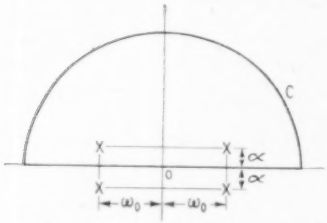

Fig. 2. *Contour for integration in ω-plane: poles of integrand marked by X.*

The time waveform $f(t)$ has an amplitude spectrum $g(\omega)$, where ω equals 2π times frequency, given by the Fourier integral[7]

$$g(\omega) = \frac{1}{2\pi} \int_{-\infty}^{\infty} f(t) e^{-j\omega t}\, dt \quad .. \quad .. \quad .. \quad (5)$$

The amplitude spectrum for the waveform $F(t)$ is therefore

$$G(\omega) = \frac{1}{2\pi} \int_{-\infty}^{\infty} f(t) e^{-j\omega t}\, dt + \frac{1}{2\pi} \int_{-\infty}^{\infty} f(t-\tau) e^{-j\omega t}\, dt$$
$$= g(\omega) + e^{-j\omega\tau}\, g(\omega) \quad .. \quad .. \quad .. \quad (6)$$

The detector output may be expressed in terms of $G(\omega)$ by Parseval's theorem which states[8]

$$\int_0^{\infty} |G(\omega)|^2\, d\omega = \frac{1}{4\pi} \int_{-\infty}^{\infty} |F(t)|^2\, dt \quad .. \quad (7)$$

and so $$S = 4\pi K \int_0^{\infty} |G(\omega)|^2\, d\omega \quad .. \quad .. \quad (8)$$

The expression $|G(\omega)|^2\, d\omega$ represents the incident power at the detector which lies in the range of angular frequencies from ω to $\omega + d\omega$. It is possible at this stage to make allowance for the inevitable frequency dependence of the detector response. Suppose that the detector response is $h(\omega)$ for unit incident power at the angular frequency ω. Then integrating over all angular frequencies gives for the detector response in the present case

$$S = \int_0^{\infty} h(\omega)\, |G(\omega)|^2\, d\omega \quad .. \quad .. \quad .. \quad (9)$$

The constant $4\pi K$ of the previous equation has been included in the function $h(\omega)$. Now, from equation (6)

$$|G(\omega)|^2 = |g(\omega)|^2\, |1 + e^{-j\omega\tau}|^2$$
$$= 2\, |g(\omega)|^2\, (1 + \cos \omega\tau) \quad .. \quad (10)$$

Hence,

$$S = 2 \int_0^{\infty} h(\omega) |g(\omega)|^2\, d\omega +$$
$$2 \int_0^{\infty} h(\omega) |g(\omega)|^2 \cos \omega\tau\, d\omega \quad .. \quad .. \quad (11$$

which gives the detector output as a function of τ. The right-hand side consists of two terms which are equal when τ is zero. The second, which depends on τ, may be written $P(\tau)$ and tends to zero as τ tends to infinity for the waveforms of practical interest. The expression for $P(\tau)$ is

$$2 \int_0^{\infty} h(\omega)\, |g(\omega)|^2 \cos \omega\tau d\omega$$

i.e., the Fourier cosine transform of $h(\omega)|g(\omega)|^2$. By the inverse transform,

$$h(\omega)\, |g(\omega)|^2 = \frac{1}{\pi} \int_0^{\infty} P(\tau) \cos \omega\tau\, d\omega \quad .. \quad (12)$$

The use of this equation depends on extracting the function $P(\tau)$ from the experimentally-obtained interferogram. The relation between τ and s is given immediately by equation (2) and it is then only necessary to subtract from the interferogram the constant term corresponding to the first integral in equation (11). The value of this constant equals the value of S for very large s, since $P(\tau)$ tends to zero as τ tends to infinity. If the apparatus has been correctly aligned, the constant should also equal half the value, which S has when s is zero. When $P(\tau)$ has been obtained in this way, the integration on the right-hand side of equation (12) may be carried out, using if necessary numerical or mechanical methods, to give the function $h(\omega)\, |g(\omega)|^2$. If $h(\omega)$, the detector response, is known, the power spectrum of the source $|g(\omega)|^2$ is immediately obtained.

A simple example of the foregoing analysis is provided by the time waveform,

$$f(t) = 0 \qquad t < 0$$
$$= e^{-\alpha t} \sin \omega_0 t \quad t \geqslant 0 \quad .. \quad .. \quad (13)$$

which corresponds fairly closely to the damped wavetrain emitted by a sparked oscillator. The resonant frequency of the sparked resonator is $\omega_0/2\pi$ and the damping constant is α. The amplitude spectrum, calculated from equation (5), is

$$g(\omega) = -\frac{\omega_0}{2\pi(\omega - \omega_0 - j\alpha)(\omega + \omega_0 - j\alpha)} \quad (14)$$

and so $$g(\omega)|^2 = \frac{{\omega_0}^2}{4\pi^2(\omega - \omega_0 - j\alpha)(\omega + \omega_0 - j\alpha)(\omega - \omega_0 + j\alpha)(\omega + \omega_0 + j\alpha)} \quad .. \quad .. \quad .. \quad (15)$$

It will be assumed for simplicity that $h(\omega)$ is a constant (i.e., that the detector has a uniform frequency response) and so from equation (11),

$$P(\tau) = \frac{{\omega_0}^2}{2\pi^2}\int_0^\infty \frac{\cos \omega\tau}{(\omega - \omega_0 - j\alpha)(\omega + \omega_0 - j\alpha)(\omega - \omega_0 + j\alpha)(\omega + \omega_0 + j\alpha)} d\omega \quad .. \quad .. \quad (16)$$

Since the integrand is an even function of ω, this may be written

$$P(\tau) = \frac{{\omega_0}^2}{4\pi^2}\int_{-\infty}^\infty \frac{\exp(j\omega\tau)}{(\omega - \omega_0 - j\alpha)(\omega + \omega_0 - j\alpha)(\omega - \omega_0 + j\alpha)(\omega + \omega_0 + j\alpha)} d\omega \quad .. \quad (17)$$

The integral can be most easily evaluated by applying the Cauchy residue theorem[9] to the contour shown in Fig. 2, C being a semicircle whose radius is allowed to tend to infinity. The integral round C then vanishes, provided τ is positive, and so

$$P(\tau) = \frac{{\omega_0}^2}{4\pi^2} \times 2\pi j \times (\text{sum of the residues at the poles of the integrand in the upper half-plane of Fig. 2})$$

$$= \frac{j{\omega_0}^2}{2\pi}\left[\frac{\exp j(\omega_0 + j\alpha)\tau}{8j\omega_0\alpha(\omega_0 + j\alpha)} + \frac{\exp -j(\omega_0 - j\alpha)\tau}{8j\omega_0\alpha(\omega_0 - j\alpha)}\right]$$

$$= P(0)\sec\phi\, e^{-\alpha\tau}\cos(\omega_0\tau - \phi) \quad .. \quad (18)$$

where $P(0) = {\omega_0}^2/[8\pi\alpha({\omega_0}^2 + \alpha^2)]$

and $\phi = \tan^{-1}(\alpha/\omega_0)$

From the results given previously on the relation between $P(\tau)$ and the interferogram, it is found that the interferogram for the present case is given by

$$S = \tfrac{1}{2}S_0[1 + \sec\phi.\exp(-2\alpha s/\tau).\cos(2\omega_0 s/c - \phi)] \quad .. \quad .. \quad (19)$$

where S_0 is the value of S when s is zero. The general characteristics of the time waveform, the power spectrum and the interferogram are shown by the sketch of Fig. 3. In particular the interferogram has the appearance of a damped sinusoidal wave, the period of the oscillation being half the free space wavelength of the radiation; i.e., $\pi c/\omega_0$.

It may be noted that the part of the interferogram defined by $P(\tau)$ is directly proportional to the auto-correlation function of the time waveform.[10] The result in equation (12) is therefore identical to that frequently used in the study of noise problems, where it is known as Wiener-Khintchine theorem.[11]

It has been shown that the power spectrum of a source can be obtained completely from an interferogram measured using a Boltzmann interferometer. This result should be of value in investigations of the fundamental mechanism of sparked resonator oscillators.

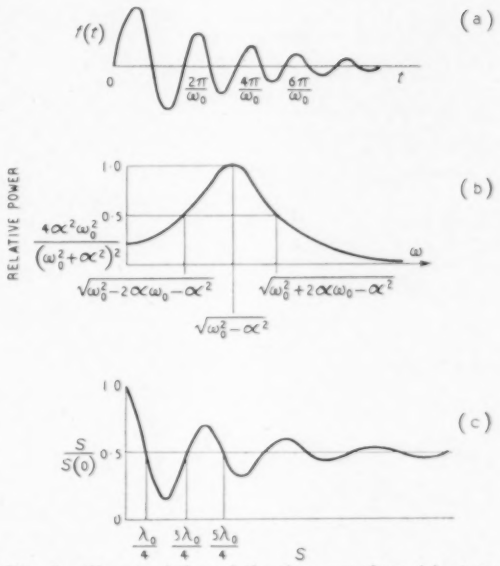

Fig. 3. *Characteristics of the time waveform (a), power spectrum (b) and the interferogram (c) for the example discussed in the text.*

Since this paper was prepared, the authors have learned that Mr. M. H. N. Potok, of the Royal Technical College, Glasgow, is engaged in similar work and hopes to publish examples of the method as applied to the observed radiation from sparked oscillators.

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BANDPASS AMPLIFIERS

Investigation of Design and Stability

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SUMMARY.—In this paper an approach to bandpass amplifier design is presented. This is based on the required overall gain, half-power bandwidth and centre frequency of the pass-band. The amplifier considered is composed of a number of identical groups, each consisting of one or several valves and circuits, having a maximum-flatness frequency response. The conditions for maximum possible gain are found and a graph is given which makes it possible to calculate the number of groups necessary to fulfil the requirements.

The effect of anode–grid capacitances on the performance of single-tuned and stagger-tuned amplifiers is discussed. It is shown that the amount of distortion introduced into the frequency–amplitude response curves depends on a regeneration coefficient α. When α is sufficiently large, oscillations occur. A single-tuned amplifier of any number of stages is inherently stable if $\alpha < \frac{1}{2}$. It is shown that, with stagger-tuned amplifiers, the amount of distortion depends not only on the magnitude of α, but also on the order of staggering. Values of α necessary to produce oscillations in a staggered pair, staggered triple and m staggered pairs are given. The effect of the regeneration on the process of tuning the amplifiers is also discussed.

1. Introduction

MOST of the bandpass amplifiers, especially those operating linearly, can be designed by using appropriate formulae. In general, the design is based on the required gain and bandwidth. The number of stages necessary to satisfy these requirements may be evaluated from a transcendental equation. Each type of amplifier, however, requires the solution of a different equation. In this paper an attempt will be made to present a single approach, which will be valid for several types of bandpass amplifiers. The conditions of maximum obtainable gain will also be derived.

It seems also that the question of bandpass amplifier stability has not been fully worked out. One type of regeneration inherent in bandpass amplifiers is due to anode–grid capacitances. This effect is well known and the criteria of stability have been obtained for certain cases. A more thorough approach to this problem will, however, be attempted: the shape of frequency–amplitude response curves, under conditions of appreciable regeneration, will be investigated, and the conditions of stability for single-tuned and stagger-tuned amplifiers will be given. Also it will be shown how the criteria of stability affect the design of a bandpass amplifier.

The treatment given in this paper is purely theoretical. It should be added, however, that a special five-stage amplifier was constructed for the purpose of experimental investigation. The experimental results, though not highly accurate, seem to corroborate the theory which is here presented. The amplifier employed EF91 miniature pentodes and operated at a frequency of 10 Mc/s. Grid–anode capacitances were artificially increased, by small capacitors, up to about 0·25 pF per valve.

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2. An Approach to Bandpass Amplifier Design

The design of a bandpass amplifier is often based on the following data: the required overall gain, centre frequency and half-power bandwidth. The above data will be taken as the basis for calculations and it will be assumed that the mutual conductances of valves and the tuning capacitances of the circuits are all equal. A bandpass amplifier composed of n identical groups will be considered. Each group may consist of one or more amplifying stages and is characterized by maximum flatness amplitude response given by

$$K_m = \frac{K_{0m}}{(1 + x^{2m})^{\frac{1}{2}}} \qquad \ldots \qquad \ldots \qquad (1)$$

where:

K_{0m} = maximum gain of a group

K_m = gain of a group as a function of frequency

m = number of tuned circuits in a group

$x = \dfrac{f_0}{B_m}\left(\dfrac{f}{f_0} - \dfrac{f_0}{f}\right)$ or approximately $x \approx \dfrac{2\Delta f}{B_m}$

f = frequency

f_0 = centre frequency

Δf = detuning from the centre frequency

B_m = overall bandwidth of a group at points 3 db down.

The following types of groups will be considered:—

(a) one valve with one single-tuned circuit;

(b) one valve with a double-tuned, critically-coupled circuit;

(c) m valves with m staggered single-tuned circuits;

(d) group employing double-tuned coupled circuits with, or without, one single-tuned circuit. All the circuits are separated by valves.

If m is even, the maximum flatness response curve may be obtained by using double-tuned circuits of appropriate Q values and coupling coefficients. If m is odd, an additional single-tuned circuit should be employed.

The design and the characteristics of these basic amplifying units have been investigated by various authors; in particular, much information on the subject can be found in the works of D. Weighton,[1] H. Wallman[2] and A. G. W. Uitjens.[3] It is felt, however, that some explanation should be given with regard to group (d); the analysis of this amplifying unit is given in Appendix 1.

The amplitude response of an amplifier composed of n groups is

$$[K_m]_n = \frac{K_{0m}{}^n}{(1 + x^{2m})^{n/2}} \quad .. \quad .. \quad (2)$$

The overall half-power bandwidth of the amplifier is related to the bandwidth of a group in the following way :—

$$\left[1 + \left(\frac{2\Delta f_1}{B_m}\right)^{2m}\right]^{n/2} = \sqrt{2} \quad .. \quad .. \quad (3)$$

where

$$2\Delta f_1 = B_n$$

Thus

$$B_n = B_m(2^{1/n} - 1)^{1/2m} \quad .. \quad .. \quad (4)$$

The maximum overall gain of the amplifier consisting of n groups may be expressed in terms of the overall bandwidth B_n, the tuning capacitances C and the mutual conductances g_m of the valves (assumed to be equal for all circuits).

Maximum gain of a group is

$$K_{0m} = \left(\frac{g_m}{2\pi CB_m}\right)^t \times s \quad .. \quad .. \quad (5)$$

where

t = the number of valves in a group

s = a coefficient depending on the type of a group. It will be shown later that $s \leqslant 1$

It follows from (4) and (5) that the maximum gain of n groups is

$$[K_{0m}]_n = K_{0m}{}^n = \left(\frac{g_m}{2\pi CB_m}\right)^{tn} \times s^n$$

$$= \left(\frac{g_m}{2\pi CB_n}\right)^{tn} \times s^n \times (2^{1/n} - 1)^{tn/2m} \quad (6)$$

Let $$b = \frac{g_m}{2\pi CB_n}$$

Then

$$[K_{0m}]_n = b^{tn} \times s^n \times (2^{1/n} - 1)^{tn/2m} \quad (7)$$

It should be noticed that amplification is possible only when $b > 1$, since $s \leqslant 1$.

The number of valves employed in the amplifier is

$$N = nt \quad .. \quad .. \quad .. \quad .. \quad (8)$$

The maximum gain may be also expressed by

$$[K_{0m}]_n \approx b^{tn} \times s^n \times \left(\frac{\log_e 2}{n}\right)^{tn/2m} \quad .. \quad (9)$$

since $$2^{1/n} - 1 \approx \log_e 2/n \quad .. \quad .. \quad (10)$$

This latter approximation is quite close, provided $n > 2$.

Taking decimal logarithms on both sides of Equ. (9), one obtains

$$\log [K_{0m}]_n = \frac{nt}{2m} \times \log [b^{2m} \times s^{2m/t} \times (\log_e 2/n)] \quad .. \quad .. \quad .. \quad (11)$$

Let

$$A_m = b^{2m} \times s^{2m/t} \times (\log_e 2) \quad .. \quad .. \quad (12)$$

Then

$$\log [K_{0m}]_n = \frac{nt}{2m} \times \log (A_m/n) \quad .. \quad .. \quad (13)$$

The gain may be expressed in decibels

$$k_{mn} = 20 \log(K_{0m})_n = 10 \frac{nt}{m} \log\left(\frac{A_m}{n}\right) \quad .. \quad (14)$$

Therefore

$$m\frac{k_{mn}}{t} = 10n \log\left(\frac{A_m}{n}\right)$$

Now, let

$$F\left(\frac{n}{A_m}\right) = 10 \frac{n}{A_m} \times \log\left(\frac{A_m}{n}\right) \quad .. \quad (15)$$

Then, it follows that

$$\frac{m}{t}\frac{k_{mn}}{A_m} = 10 \frac{n}{A_m} \times \log\left(\frac{A_m}{n}\right) \quad .. \quad (16)$$

and

$$F\left(\frac{n}{A_m}\right) = \frac{m}{t} \times \frac{k_{mn}}{A_m} \quad .. \quad .. \quad (17)$$

Equ. (17) with unknown n, is transcendental, but it can be solved graphically. For this purpose the function $F(n/A_m)$ is plotted in Fig. 1, while Fig. 2 shows it on an enlarged scale up to its maximum. The function $F(n/A_m)$ has a maximum value of 1·6 at $n/A_m = 1/e$, and it is equal to zero at $n/A_m = 1$ and $n/A_m = 0$.

The four types of amplifiers mentioned above will now be discussed in some detail.

(a) *Multistage Single-Tuned Amplifiers*

The group is formed of one amplifying stage: one valve followed by a single tuned circuit. For this case $m = t = 1$; also $s = 1$ (see Appendix 1).

The response curve of the amplifier is

$$K_{1n} = \frac{[K_{01}]_n}{(1 + x^2)^{n/2}} \quad .. \quad .. \quad (18)$$

It can be seen that with a given A_1, the maximum possible amplification which can be achieved,

occurs when $F(n/A_1)$ is maximum. Therefore

$$[k_{1n}]_{max} = \frac{10A_1}{e} \log e = 1{\cdot}6A_1 \text{ db} \quad .. \quad (19)$$

where

$$A_1 = b^2 \times (\log_e 2) = \left(\frac{g_m}{2\pi CB_n}\right)^2 \times (\log_e 2)$$

At the point of maximum gain, the required number of stages is $n = A_1/e$.

Therefore

$$[k_{1n}]_{max} = 10n \log e = 4{\cdot}34n \text{ db} \quad .. \quad (20)$$

Thus, the gain per stage, when the maximum is reached, is given by

$$1/n \times [k_{1n}]_{max} = 10 \times \log e = 4{\cdot}34 \text{ db} \; .. \quad (21)$$

This result was obtained by H. Wallman[1] and W. Kleen.[2]

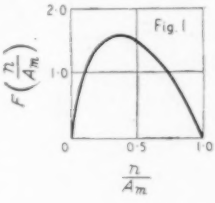
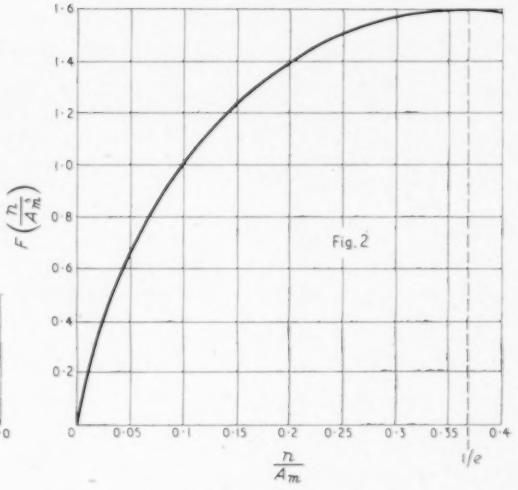


Fig. 1 (left). *Curve from which the required number of stages can be obtained.*

Fig. 2. (right) *Enlarged portion of Fig.* 1.



(b) *Double-Tuned Coupled-Circuit Amplifier*

Each group consists of one valve with a pair of tuned, critically-coupled, circuits. This type of amplifier is only a particular case of the type to be described in (d). It will be discussed in some detail, however, since this amplifier is widely used. It can be seen (from Appendix 1) that in this case $m = 2$ and $t = 1$, while $s = 1/\sqrt{2}$.

The gain of one stage is

$$K_2 = \frac{K_{02}}{(1 + x^4)^{\frac{1}{2}}} \quad .. \quad .. \quad .. \quad (22)$$

where

$$K_{02} = \frac{g_m}{2\sqrt{2}\pi CB_2}$$

B_2 is the half-power bandwidth of a single stage.

The gain of n such stages is

$$K_{2n} = \frac{[K_{02}]_n}{(1 + x^4)^{n/2}} = \frac{K^n{}_{02}}{(1 + x^4)^{n/2}} \quad (23)$$

The overall bandwidth is related to the bandwidth of a single stage in the following manner:

$$B_n = B_2(2^{1/n} - 1)^{\frac{1}{4}} \quad .. \quad .. \quad .. \quad (24)$$

Also, one obtains from (12)

$$A_2 = \left(\frac{\log_e 2}{4}\right) \times \left(\frac{g_m}{2\pi CB_n}\right)^4 \quad .. \quad (25)$$

The maximum possible gain in db is given by the expression

$$[k_{2n}]_{max} = 10 \frac{A_2 t}{em} \log e = 0{\cdot}8A_2 \text{ db}. \; .. \quad (26)$$

And the required number of stages is $n = A_2/e$. Therefore,

$$[k_{2n}]_{max} = 10nt/m \log e = 2{\cdot}17n \text{ db}.$$

(c) *Amplifier Composed of Stagger-Tuned Groups*

In this case, a group consists of m stagger-tuned stages. It has a maximum-flatness amplitude response as given by Equ. (1). There are as many valves as there are tuned circuits; thus for this case $t = m$.

The maximum gain of a group is

$$K_{0m} = \left(\frac{g_m}{2\pi CB_m}\right)^m \quad .. \quad .. \quad (27)$$

thus, it follows from (5) that $s = 1$.

As before, the maximum possible gain in db is

$$[k_{mn}]_{max} = 1{\cdot}6A_m$$

and it occurs when the number of groups is $n = A_m/e$, where

$$A_m = (\log_e 2)b^{2m} = (\log_e 2) \times \left(\frac{g_m}{2\pi CB_n}\right)^{2m}$$

It may be noticed that the maximum obtainable gain is proportional to b^{2m}, while for a single-tuned amplifier it is proportional to b^2. It is clear, then, that with the same b factors in both cases, the amplifier composed of staggered groups offers some advantages.

(d) *Amplifier Composed of Double-Tuned (plus Single-Tuned) Groups*

There are two cases. If m is even, the curve given by

$$K_m = \frac{K_{0m}}{(1 + x^{2m})^{\frac{1}{2}}}$$

can be realized by using $\frac{1}{2}m$ double-tuned circuits of appropriate coupling coefficients and dynamic impedances. There are $t = \frac{1}{2}m$ valves in a group.

For this case

$$s = 2^{-(m-1/2)}$$

The derivation of this expression is given in Appendix 1.

Thus

$$\begin{aligned} A_m &= b^{2m} \times s^{2m/t} \times (\log_e 2) \\ &= b^{2m} \times s^4 \times (\log_e 2) \\ &= (g_m/2\pi CB_n)^{2m} \times 2^{-2(m-1)} \times (\log_e 2) \quad (28) \end{aligned}$$

If m is odd, $\frac{1}{2}(m-1)$ double-tuned pairs and one single-tuned stage unit must be employed in order to obtain the desired amplitude response for a group. Thus, the number of valves per group is $t = \frac{1}{2}(m+1)$, and the total number of valves in the amplifier is given by

$$N = \tfrac{1}{2}(m+1)n \quad .. \quad .. \quad .. \quad .. \quad (29)$$

In this case

$$s = \frac{\sqrt{m}}{2^{(m-1)/2}}$$

Therefore

$$\begin{aligned} A_m &= b^{2m} \times s^{2m/t} \times (\log_e 2) \\ &= b^{2m} \times s^{4m/m+1} \times (\log_e 2) \\ &= \left(\frac{g_m}{2\pi CB_n}\right)^{2m} \times \left(\frac{\sqrt{m}}{2^{(m-1/2)}}\right)^{4m/m+1} \times (\log_e 2) \\ &\quad .. \quad .. \quad .. \quad (30) \end{aligned}$$

Again, the maximum possible gain is

$$[k_{mn}]_{max} = 1{\cdot}6 \times \frac{t}{m} \times A_m \text{ db}$$

and the number of groups required is $n = A_m/e$.

(e) *Design Procedure*

In general, $b = g_m/2\pi CB_n$ is appreciably greater than unity. Thus, by employing a suitable type of amplifier it is possible to obtain any desired amplification. The required gain, however, will probably never be higher than 120 db.

The procedure in the design of an amplifier is as follows: first, b should be calculated. For this purpose, the mutual conductances of the valves, the tuning capacitances of the circuits and the overall bandwidth should be known. Then, the type of group should be selected. Also, it must be decided whether stagger-tuned groups, or groups employing coupled circuits, are preferred.

Once that is done, the value of $F\left(\frac{n}{A_m}\right) = \frac{m}{t} \times \left(\frac{k_{mn}}{A_m}\right)$ is computed. The corresponding point can be found on the curve in Fig. 2. Thus, the number of groups n is obtained. For a given amplification and bandwidth, several types of amplifiers may be tried and the preference given to the one which is most convenient and which requires the least number of valves.

It can be seen from (6) that the gain is proportional to the factor $g_m/2\pi C$. Thus, it seems to be desirable to make the factor $g_m/2\pi C$ as high as possible in order to obtain the desired amplification with the minimum number of stages. For a given mutual conductance g_m, a high value of $g_m/2\pi C$ can be obtained by employing small tuning capacitances C. The choice of C, however, is not quite arbitrary; if C is small (which implies high dynamic impedances), the amplifier may become unstable owing to the regenerative effect of grid-to-anode capacitances. On the other hand, even if the amplifier is stable, it may work under conditions of fairly high regeneration and then the analysis which was given above is not valid. Thus, it is necessary to select such values of C as to obtain a stable amplifier working under conditions of negligible regeneration. This problem is discussed in the following section.

3. Stability of Bandpass Amplifiers.

It is generally understood that the grid-to-anode capacitances C_{ga} cause regeneration or positive feedback in a multi-stage bandpass amplifier and the problem has been considered by various authors. M. O'Connor Horgan[5] thoroughly investigated the properties of a single amplifying stage with two tuned circuits, and gave the oscillation conditions for a single-tuned amplifier of up to four stages. The criteria of oscillation in a single-tuned amplifying chain were also examined by W. Faust and H. M. Beck[6]. Recently, A. G. W. Uitjens[3] presented a general but approximate method for the evaluation of permissible stable gain in stagger-tuned and coupled-circuit amplifiers. In this section it is proposed to carry the problem a little further.

(a) *Fundamental Equation*

A chain amplifier employing n valves will be considered. No assumptions will be made as to the anode impedances, except that they are two-terminal networks. The internal impedance of the generator is assumed to be $Z_{in} = 0$. The grid-to-anode impedance of the kth stage, due to C_{ga} which provides a feedback path, will be denoted by $\mathbf{Z}_{fk}$. All the complex quantities will be denoted by black-face letters; the moduli of complex quantities will be represented by ordinary letters. The equivalent circuit of a chain amplifier may be represented as shown in Fig. 3.

The anode impedance $\mathbf{Z}_k$ takes into account the output capacitance of the preceding valve and the input capacitance of the following valve. Thus $\frac{1}{\mathbf{Z}_k} = \frac{1}{r_{ak}} + \frac{1}{\mathbf{Z}_{k'}} + j\omega(C_{gk} + C_{ak})$

where

$\mathbf{Z}_{k'}$ = the impedance connected between the two valves,

r_{ak} = the internal resistance of the kth valve.

Denote

$$g_{mk} - \frac{1}{\mathbf{Z}_{fk}} = \mathbf{g}_{mk}$$

and

$$\frac{1}{\mathbf{Z}_k} + \frac{1}{\mathbf{Z}_{fk}} + \frac{1}{\mathbf{Z}_{f(k+1)}} = \frac{1}{\mathbf{Z}_{kk}}$$

Then, the system of equations, relating the voltages at various stages of the amplifying chain, may be written in the matrix form as shown below:

$$\begin{bmatrix} \frac{1}{\mathbf{Z}_{11}} & \frac{-1}{\mathbf{Z}_{f2}} & 0 & 0 & . & 0 & 0 \\ \mathbf{g}_{m2} & \frac{1}{\mathbf{Z}_{22}} & \frac{-1}{\mathbf{Z}_{f3}} & 0 & . & 0 & 0 \\ 0 & \mathbf{g}_{m3} & \frac{1}{\mathbf{Z}_{33}} & \frac{-1}{\mathbf{Z}_{f4}} & . & . & . \\ . & . & . & . & . & . & . \\ 0 & 0 & . & . & \mathbf{g}_{m,\,n-1} & \frac{1}{\mathbf{Z}_{n-1,\,n-1}} & \frac{-1}{\mathbf{Z}_{fn}} \\ 0 & 0 & . & . & 0 & \mathbf{g}_{mn} & \frac{1}{\mathbf{Z}_{nn}} \end{bmatrix} \begin{bmatrix} \mathbf{V}_1 \\ \mathbf{V}_2 \\ . \\ . \\ \mathbf{V}_{n-1} \\ \mathbf{V}_n \end{bmatrix} = \begin{bmatrix} -\mathbf{g}_{m1}\,\mathbf{V}_g \\ 0 \\ . \\ . \\ 0 \\ 0 \end{bmatrix} \qquad . . \quad . . \quad (31)$$

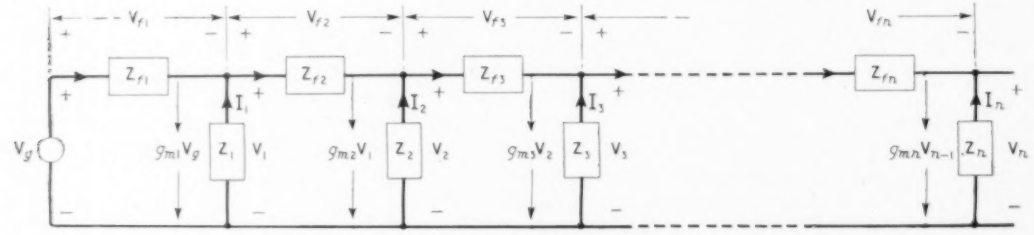


Fig. 3. *Equivalent circuit of amplifier.*

It can be deduced that the overall amplification of the chain may be expressed as

$$\mathbf{K}_n = \frac{\mathbf{V}_n}{\mathbf{V}_g} = \frac{(-1)^n \prod_{k=1}^{n} \mathbf{g}_{mk}}{\mathbf{D}_n} \qquad (32)$$

where $\mathbf{D}_n$ is the determinant of the network matrix.

In general, the determinant is a complex quantity

$$\mathbf{D}_n = A_n + jB_n$$

The system will become unstable when the determinant is equal to zero which implies that

$$A_n = B_n = 0$$

General expressions for gain up to four stages are given in Table 1.

(b) *Single-Tuned Amplifiers*

For the purpose of the investigation of single-tuned amplifiers a number of assumptions and simplifications will be made, without which the calculations would be practically impossible. Thus, it will be assumed that all the mutual conductances of valves, the anode-to-grid capacitances, and the anode impedances are

TABLE 1

n	$\mathbf{K}_n$
1	$-\mathbf{g}_{m1}\,\mathbf{Z}_{11}$
2	$\dfrac{\mathbf{g}_{m1}\,\mathbf{g}_{m2}}{\frac{1}{\mathbf{Z}_{11}\mathbf{Z}_{22}} + \frac{\mathbf{g}_{m2}}{\mathbf{Z}_{f2}}}$
3	$\dfrac{-\mathbf{g}_{m1}\,\mathbf{g}_{m2}\,\mathbf{g}_{m3}}{\frac{1}{\mathbf{Z}_{11}\mathbf{Z}_{22}\mathbf{Z}_{33}} + \frac{\mathbf{g}_{m2}}{\mathbf{Z}_{f2}\mathbf{Z}_{33}} + \frac{\mathbf{g}_{m3}}{\mathbf{Z}_{f3}\mathbf{Z}_{11}}}$
4	$\dfrac{\mathbf{g}_{m1}\,\mathbf{g}_{m2}\,\mathbf{g}_{m3}\,\mathbf{g}_{m4}}{\frac{1}{\mathbf{Z}_{11}\mathbf{Z}_{22}\mathbf{Z}_{33}\mathbf{Z}_{44}} + \frac{\mathbf{g}_{m2}}{\mathbf{Z}_{33}\mathbf{Z}_{44}\mathbf{Z}_{f2}} + \frac{\mathbf{g}_{m3}}{\mathbf{Z}_{11}\mathbf{Z}_{44}\mathbf{Z}_{f3}} + \frac{\mathbf{g}_{m4}}{\mathbf{Z}_{11}\mathbf{Z}_{22}\mathbf{Z}_{f4}} + \frac{\mathbf{g}_{m2}\,\mathbf{g}_{m4}}{\mathbf{Z}_{f2}\mathbf{Z}_{f4}}}$

identical. Furthermore, Q values of the anode tuned circuits are to be of the order of 10 or more.

Also, $\mathbf{g}_{mk} = g_{mk} - j\omega C_{gak} \approx g_{mk} = g_m$ since, in general, $g_m \gg \omega C_{ga}$.

Admittances forming the diagonal of the determinant may be expressed:

$$\frac{1}{\mathbf{Z}_{kk}} = \frac{1}{\mathbf{Z}_k} + j2\omega C_{ga} \approx \frac{1 + jx}{r_d} \quad .. \quad (33)$$

where

$$x = Q_0(f/f_0 - f_0/f) \approx \frac{2Q_0\Delta f}{f_0} = \frac{2\Delta f}{B'} \quad .. \quad (34)$$

where B' = the bandwidth of a single circuit.

$r_d = \dfrac{Q_0}{\omega_0 C}$ = the dynamic impedance of a circuit.

After a number of algebraic operations the determinant of the network may be written in the form:

$$\mathbf{D}_n = \frac{1}{{r_d}^n} \times \begin{vmatrix} 1 + jx & -j\alpha & 0 & 0 & . & 0 & 0 \\ 1 & 1 + jx & -j\alpha & 0 & . & 0 & 0 \\ 0 & 1 & 1 + jx & -j\alpha & . & . & . \\ . & . & . & . & . & . & . \\ 0 & 0 & . & . & 1 & 1 + jx & -j\alpha \\ 0 & 0 & . & . & 0 & 1 & 1 + jx \end{vmatrix} \quad (35)$$

where $\alpha = \omega C_{ga} g_m {r_d}^2 \approx \omega_0 C_{ga} g_m {r_d}^2$; it will be called the *regeneration coefficient*. It will be shown that the amount of regeneration and, ultimately, the oscillations, depend on the magnitude of α. The regeneration coefficient at which the oscillations result will be called the critical regeneration coefficient and will be denoted by α_n, where n is the number of stages in the given amplifier.

The oscillation conditions are: $\mathbf{D}_n = 0$, or $\mathbf{D}_{an} = {r_d}^n \mathbf{D}_n = 0$.

If $\mathbf{D}_{an} = A_{an} + jB_{an}$, then

$$A_{an} = B_{an} = 0 \quad .. \quad .. \quad .. \quad .. \quad (36)$$

These are two simultaneous equations whose solutions will give the values of x and α_n at the points of instability. In general, there may be more than one solution for x, but only negative x are allowed, since the equivalent input impedance of an amplifier (of any number of stages) is always composed of resistance and capacitance in parallel. This means that the first tuned circuit must appear as a resistance and inductance in parallel. The oscillations will then invariably take place at a frequency lower than the frequency to which the circuits are tuned. On the other hand, only the lowest real positive values of α_n are of interest, because these are minimum values necessary to cause instability.

A heuristic proof is given in the Appendix 2, which shows that $x = -1$, is a solution for any n. Upon its substitution into $\mathbf{D}_{an}$, an equation in terms of α only is obtained, from which the critical regeneration coefficient α_n can be computed.

The equation whose solutions yield critical regeneration coefficients is

$$(n+1)\left(\frac{1}{2}\right)^{\frac{n}{2}} + \binom{n+1}{3}\left(\frac{1}{2}\right)^{\frac{n-2}{2}}\left(\frac{1}{2} - \alpha\right) + \binom{(n+1)}{5}\left(\frac{1}{2}\right)^{\frac{n-4}{2}}\left(\frac{1}{2} - \alpha\right)^2 + \ldots = 0 \quad (37)$$

The solutions of this equation are given in Table 2.

TABLE 2

Number of Stages	2	3	4	5	6	..	∞
α_n	2	1	0·76	$\frac{2}{3}$	0·62	..	$\frac{1}{2}$

The values for α_n agree with those obtained by O'Connor Horgan[5] and by Faust and Beck[6], except that in their n-stage amplifier they assumed $n+1$ tuned circuits and not n as in the present notation. Faust and Beck calculated the oscillation conditions up to six stages (seven in this case).

It is interesting to notice that the oscillations will not start even with a very large number of stages provided $\alpha < \frac{1}{2}$. It may also be added that a single-tuned amplifier (with all the circuits tuned alike) is the most critical case from the point of view of stability. If the circuits were detuned with respect to one another, it would be more difficult to cause oscillations; i.e., a higher value of α_n would be required. Unfortunately, the authors did not find it possible to prove conclusively that this is the case; it will be shown later, however, that in an amplifier composed of staggered pairs the value of α_n necessary to cause oscillations is higher than in a single-tuned amplifier with the same number of stages.

A single-tuned amplifier works as a regenerative amplifier when α is smaller than α_n. The amount of positive feedback and, hence, the bandwidth and maximum gain, depend on the magnitude of α, which, in turn, is proportional to C_{ga}. The influence of C_{ga} on the shape of response curves is illustrated in Fig. 4* and Fig. 5. A series of curves, with α as a parameter, are plotted for a two-stage and a three-stage amplifier (see Appendix 2). It can be seen that the curves

* In his book, A. G. W. Uitjens[3] shows a double-peaked response curve, for $\alpha = 1$, for an amplifying stage with two synchronously tuned circuits. The authors found that for the above case the response curve is single-peaked; see Fig. 4.

become asymmetrical when α is greater than zero. With an increase of α the maximum gain of the amplifier increases, while the bandwidth is reduced. Furthermore, with feedback, the maximum gain no longer occurs at the frequency to which the circuits are tuned, but below it. It can be seen from the curves that, as α tends to its critical value, the point of maximum amplification approaches $x = -1$, where the oscillations will start.

It was also found that the amplitude response curves of four- and five-stage amplifiers have only one maximum and are much the same in shape as those for two- and three-stage amplifiers. Thus, it may be expected that this is the case for any number of stages.

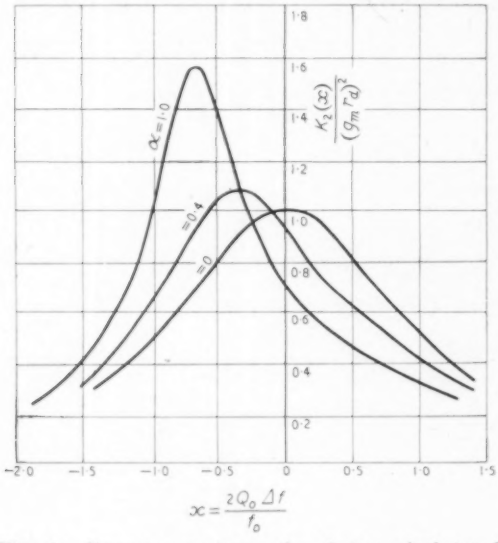


Fig. 4. *Frequency response of a 2-stage single-tuned amplifier.*

It is interesting to know how the bandwidth and the maximum gain depend on α. This is illustrated in Figs. 6 and 7 where the following functions are plotted:—

$$h_2(\alpha) = \frac{B_2(\alpha)}{B_2(0)}; \quad h_3(\alpha) = \frac{B_3(\alpha)}{B_3(0)} \quad .. \quad .. \quad (38)$$

$$g_2(\alpha) = \frac{K_{2max}}{K_{20}}; \quad g_3(\alpha) = \frac{K_{3max}}{K_{30}} \quad .. \quad .. \quad (39)$$

$B_2(0)$ is the bandwidth of a two-stage amplifier under conditions of no regeneration.

$B_2(\alpha)$ is the overall bandwidth of two stages when α is greater than zero.

K_{2max} = maximum gain when $\alpha > 0$,

$K_{20} = (g_m r_d)^2$ $\quad (\alpha = 0)$,

and so on.

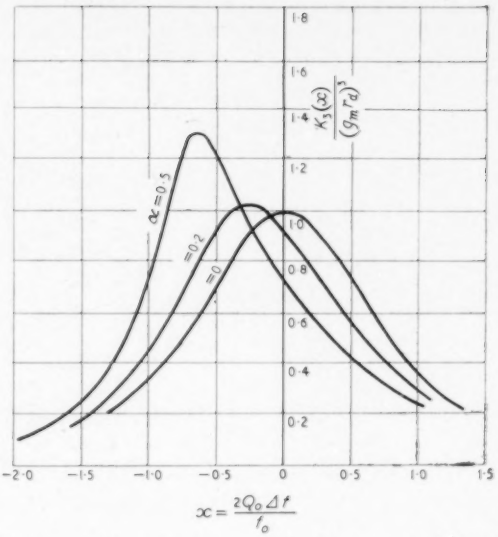


Fig. 5. *Frequency response of a 3-stage single-tuned amplifier.*

In some textbooks and papers the condition of oscillations is given in the form

$$K_0 = g_m r_d = \sqrt{\frac{\alpha_n g_m}{\omega_0 C_{ga}}} \quad .. \quad .. \quad .. \quad (40)$$

Here K_0 should be understood as the limiting value of the product $g_m r_d$ and not the maximum stable gain per stage. It may be pointed out that when regeneration is appreciable the amplification varies from stage to stage and it may be much greater than $g_m r_d$. It is only when $\alpha = 0$ that the product $g_m r_d$ does represent maximum gain per stage.

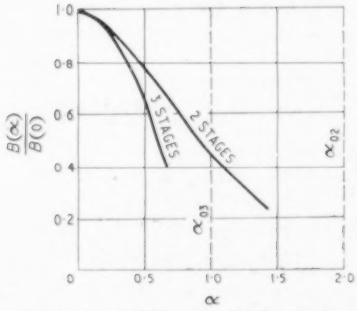


Fig. 6. *3-db bandwidth in single-tuned amplifiers.*

(c) *Stagger-Tuned Amplifiers*

It is now proposed to investigate the influence of C_{ga} on the performance of stagger-tuned amplifiers with the view to obtaining the oscillation conditions and discussing the amplitude

response curves. It will be assumed that the circuits are tuned to the desired frequencies which are computed by means of the theoretical formulae. It will be shown later how this tuning can be done. Unfortunately, no general oscillation conditions have been found, since the mathematics in the case of more than three staggered stages becomes very involved. Only the case of arithmetic symmetry of staggering will be considered; i.e., when the overall bandwidth of the amplifier is not an appreciable fraction of the centre frequency (say $B_n \leqslant 0{\cdot}2f_0$).

A *staggered pair* consists of two circuits having equal bandwidths and dynamic impedances. Its amplitude response, under the conditions of negligible regeneration, is

$$K_2 = \frac{K_{02}}{(1+x^4)^{\frac{1}{2}}} \quad .. \quad .. \quad .. \quad .. \quad (41)$$

where

$K_{02} = \frac{1}{2}(g_m r_d)^2$; $x = 2\Delta f/B_2$

B_2 is the overall bandwidth of the two-stage amplifier.

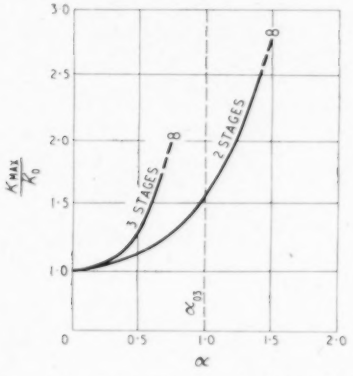


Fig. 7. *Maximum gain in single-tuned amplifiers.*

The bandwidths of single circuits are

$$B' = B_2 \sin\left(\frac{\pi}{4}\right) = B_2/\sqrt{2}$$

and the detuning from the centre frequency

$$\Delta f_1 = \pm \tfrac{1}{2}B_2 \cos\left(\frac{\pi}{4}\right) = \pm B_2/2\sqrt{2}$$

The determinant of the network is

$$\mathbf{D}_2 = \frac{1}{\mathbf{Z}_{11}\mathbf{Z}_{22}} + \frac{g_{m2}}{\mathbf{Z}_{f2}} \quad .. \quad .. \quad .. \quad (42)$$

It can be seen that it is immaterial whether the first circuit is tuned below the centre frequency and the second above it, or vice versa, because the product $1/\mathbf{Z}_{11}\mathbf{Z}_{22}$ is the same, provided the dynamic impedances and the bandwidths of both circuits are equal.

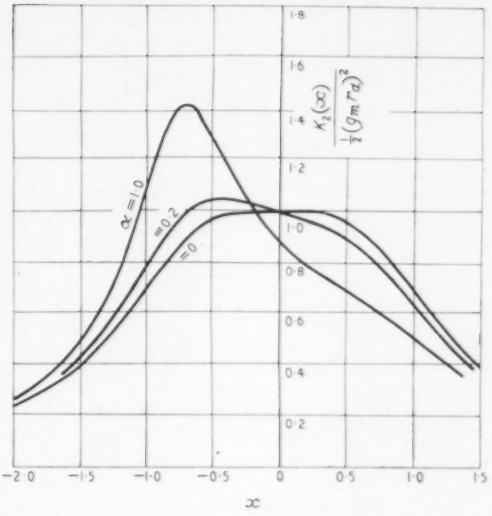


Fig. 8. *Frequency response in a 2-stage staggered amplifier.*

A series of response curves $K_2(x)$ with α as a parameter is plotted in Fig. 8. From these curves it is clear that even a small α seriously affects the shape of the characteristic. It can be seen that $\alpha = 0{\cdot}2$ already introduces a fair amount of distortion and it is probably the maximum regeneration that can be normally permitted.

In Fig. 9 the function

$$g_2(\alpha) = \frac{K_{max}}{K_{02}} \quad .. \quad .. \quad .. \quad .. \quad (43)$$

is plotted.

The effect of α on the overall bandwidth is illustrated by means of the curve

$$h_2(\alpha) = \frac{B_2(\alpha)}{B_2(0)} \quad .. \quad .. \quad .. \quad .. \quad (44)$$

which is plotted in Fig. 10.

However, even with $\alpha > 0$ it is possible to

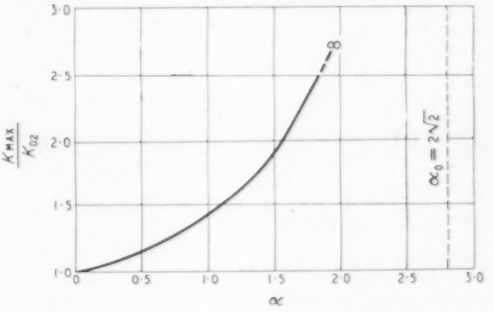


Fig. 9. *Maximum gain in a 2-stage staggered amplifier.*

obtain a symmetrical characteristic for two staggered stages by additional damping of the first circuit and the proper choice of staggering frequencies. The problem can be solved analytically without great difficulty, but the analysis is beyond the scope of this paper.

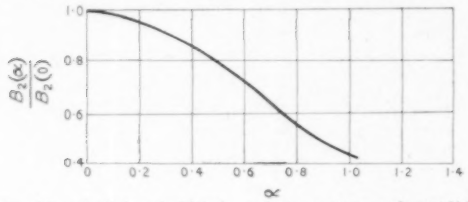


Fig. 10. *3-db bandwidth in a 2-stage staggered amplifier.*

Before an amplifier with more than two staggered stages can be discussed, it will be necessary to introduce a special notation which will represent the order in which the circuits are staggered. When the regeneration is negligible, no special significance can be attached to the order of staggering. If the feedback between the stages is appreciable, however, the amount of distortion introduced into the amplitude response curves depends on the staggering order. Furthermore, the conditions of oscillations are also functions of the way in which the circuits are staggered.

The notation is as follows:

For the kth circuit (counting from the input)

0—means that the circuit is tuned to the centre frequency

1^+—means that the circuit is tuned to the first frequency above the centre

2^+—the circuit is tuned to the second frequency above the centre

1^-—the circuit is tuned to the first frequency below the centre, and so on.

Thus, if it is written

$$S(1^+, 0, 1^-)$$

it should be understood that a three-stage amplifier has its first circuit tuned to a frequency above the centre; the second circuit is tuned to the centre frequency, and the third circuit has its resonance below the centre frequency.

Thus, for two staggered pairs there are the following possibilities:

S_1 $(1^+, 1^-, 1^+, 1^-)$
S_2 $(1^-, 1^+, 1^-, 1^+)$ } 'interlaced' staggering
S_3 $(1^-, 1^+, 1^+, 1^-)$
S_4 $(1^+, 1^-, 1^-, 1^+)$
S_5 $(1^-, 1^-, 1^+, 1^+)$
S_6 $(1^+, 1^+, 1^-, 1^-)$

Two staggering orders which give the same response characteristics and the same oscillation conditions will be called equivalent and their equivalence will be denoted $S_i \equiv S_j$.

Thus, by examining Table 1 in Section 3(a) it can be seen that, in the case of two staggered pairs, $S_2 \equiv S_1$, and $S_5 \equiv S_6$, provided the mutual conductances of all valves are equal.

In the case of 'interlaced' staggering of an amplifier consisting of m staggered pairs, a general solution can be found for oscillation conditions. It is shown in Appendix 2 that the oscillations take place at $x_1 = -1$, where $x = \sqrt{2}\,\Delta f/B'$; B' is the bandwidth of a single circuit.

The values of $\alpha = \omega_0 C_{ga} g_m r_d^2$ necessary to produce oscillations are given in Table 3.

TABLE 3

Number of Pairs m	1	2	3	..	∞
α_{2m}	$2\sqrt{2}$	$0{\cdot}76\sqrt{2}$	$0{\cdot}62\sqrt{2}$	..	$\sqrt{2}/2$

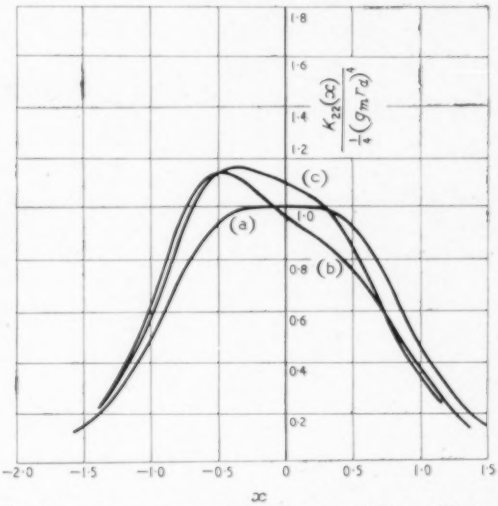


Fig. 11. *Frequency response of an amplifier with two staggered pairs of stages.*
(*a*) $\alpha = 0$
(*b*) $\alpha = 0{\cdot}2$; $S(1^+, 1^-, 1^+, 1^-)$
(*c*) $\alpha = 0{\cdot}2$; $S(1^-, 1^+, 1^+, 1^-)$

It is interesting to notice that critical α is $\sqrt{2}$ times greater than the critical regeneration coefficient in the case of a single tuned amplifier with the same number of stages and the same dynamic impedances r_d.

When the order of staggering is not 'interlaced', the conditions of oscillations, and, consequently, the response curves, are different. In order to get some insight into the matter, two staggered pairs will be considered. If equal g_m for all the

four valves are assumed, there are four non-equivalent staggering orders, namely

S_1 $(1^-, 1^+, 1^-, 1^+)$
S_3 $(1^-, 1^+, 1^+, 1^-)$
S_4 $(1^+, 1^-, 1^-, 1^+)$
S_6 $(1^+, 1^+, 1^-, 1^-)$

The oscillation conditions for those four possibilities are given in Table 4.

TABLE 4

S	S_1	S_3	S_4	S_6
x_1	− 1·0	− 0·8	− 1·27	− 0·815
α_{22}	1·073	1·34	0·96	1·18

The amplitude response curves are given in Fig. 11 for the cases S_1 and S_3. It appears from these curves that when $\alpha = 0{\cdot}2$ the distortion of characteristics is quite appreciable; thus, normally, the permissible value of α would be about 0·1 or, approximately, 10% of α_{22}.

The case of *staggered triple* may be solved analytically without great difficulty, if arithmetical symmetry of staggering is assumed.

For a three-stage amplifier the determinant of the network is

$$\mathbf{D}_3 = \frac{1}{\mathbf{Z}_{11}\mathbf{Z}_{22}\mathbf{Z}_{33}} + j\omega C_{ga} g_m \left(\frac{1}{\mathbf{Z}_{11}} + \frac{1}{\mathbf{Z}_{33}}\right) \qquad (45)$$

provided all g_m and C_{ga} terms are equal.

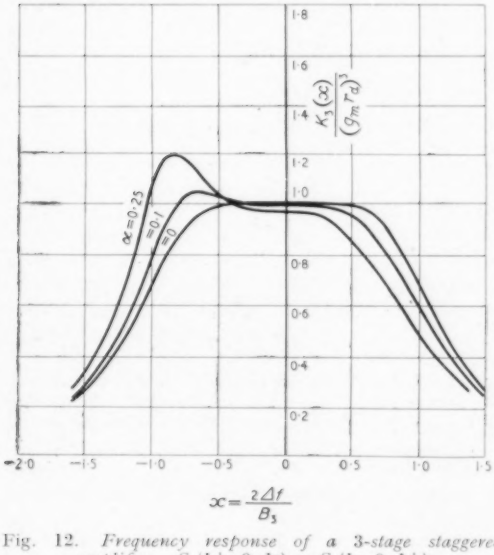


Fig. 12. *Frequency response of a 3-stage staggered amplifier.* $S_1(1^+, 0, 1^-) \equiv S_2(1^-, 0, 1^+)$

By examining the determinant it can be seen that there are three pairs of staggering orders

$$S_a(1^-, 0, 1^+) \equiv S(1^+, 0, 1^-)$$
$$S_b(0, 1^-, 1^+) \equiv S(1^+, 1^-, 0) \quad .. \quad .. \quad (46)$$
$$S_c(1^-, 1^+, 0) \equiv S(0, 1^+, 1^-)$$

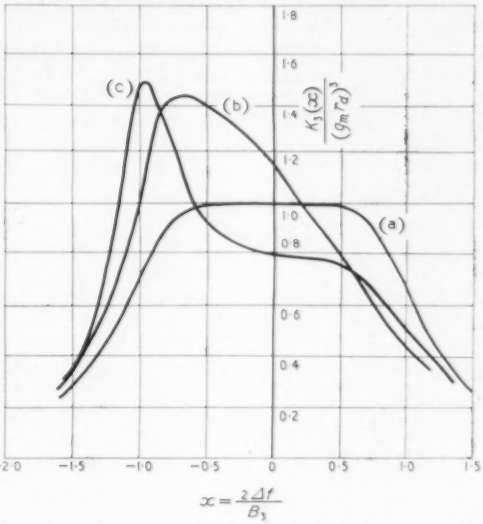


Fig. 13. *Frequency response of a 3-stage staggered amplifier.*
(*a*) $\alpha = 0$
(*b*) $\alpha = 0{\cdot}25$; $S(1^-, 1^+, 0) \equiv S(0, 1^+, 1^-)$
(*c*) $\alpha = 0{\cdot}25$; $S(1^+, 1^-, 0) \equiv S(0, 1^-, 1^+)$

The conditions of oscillations for these three cases are given in the Table 5.

TABLE 5

S	S_a	S_b	S_c
x_1	− 1·17	− 1·15	− 0·916
α_3	0·743	0·518	0·702

where

$\alpha = \omega_0 C_{ga} g_m r_d^2$
r_d = the dynamic impedance of the low Q circuit
$x = 2\Delta f / B_3$
B_3 = the bandwidth of the low Q circuit.

The effect of regeneration, due to C_{ga}, on the response characteristics is illustrated by a series of curves given in Figs. 12 and 13. For the symmetrical case S_a three curves are plotted. It can be seen that, for this case, the regeneration coefficient should not be higher than about 10% of α_3, if serious distortion of the characteristic is to be avoided. The other two cases, S_b and S_c, give more distortion and are more susceptible to oscillations than the symmetrical case.

The case of three staggered stages was investigated experimentally by M. T. Lebenbaum[7] who gave a response curve similar to that of the symmetrical case. He also found it possible to compensate for the distortion of the amplitude response curve by empirical adjustment of stagger-frequencies and damping resistances.

A staggered quadruple may be solved by the same method as that outlined above. There are $4! = 24$ different staggering orders, which, in the case of equal values of g_m, may be reduced to 12 different S_i. The analytical solution is very involved.

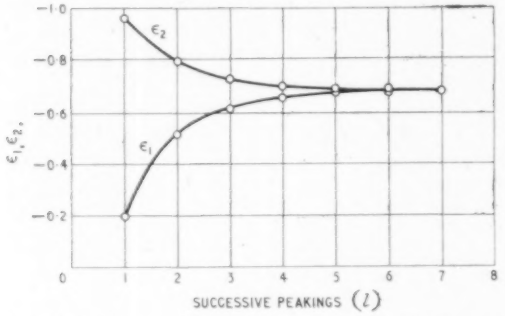


Fig. 14. *Tuning procedure in a 2-stage single-tuned amplifier* ($\alpha = 1$). *Initially* $\epsilon_2 = -2$.

(d) *Tuning*

It is of some interest to find out how the circuits of an amplifier are tuned when some regeneration, due to C_{ga}, is present.

The tuning procedure is usually as follows: an indicator is connected to the output stage, while a signal of a given frequency is applied to the input of the amplifier. In a single-tuned amplifier a single frequency f_0 is applied and all the circuits are trimmed so as to give maximum signal at the output. In a stagger-tuned amplifier, as many different frequencies are, in turn, applied as there are circuits, and for a given frequency an appropriate circuit is trimmed for maximum output at the indicator. It is assumed that the circuits are appropriately damped and have the required bandwidths. When no regeneration is present, the circuits will tune to the required frequencies, if the above procedure is adopted. This is not the case, however, when there is appreciable positive feedback between the stages. Thus, it can be seen from the curves in Figs. 4 and 5 that, if all the circuits are tuned to the same frequency f_0, maximum output occurs below f_0.

The question of tuning was approached analytically, and it was found that in a single-tuned amplifier, after a series of consecutive trimmings, the circuits did not tune to the applied frequency f_0. In a two-stage amplifier, both circuits tuned to a single frequency which was somewhat higher than f_0. The process of trimming is illustrated in Fig. 14, where ϵ_1 and ϵ_2 are the detunings from the applied frequency, of the first and second circuits respectively; ϵ_1 and ϵ_2 are expressed in terms of $x = 2\Delta f/B'$. It can be seen that, for $\alpha = 1$, after a series of peakings, the two circuits converge at $\epsilon_1 = \epsilon_2 = -0{\cdot}68$. It may be added that when $\alpha = 0$, the circuits tune to f_0 and only one trimming for each circuit is necessary.

The tuning process in a three-stage amplifier is shown in Fig. 15. It can be seen that, for $\alpha = 1/2$, the first and third stages tune to one

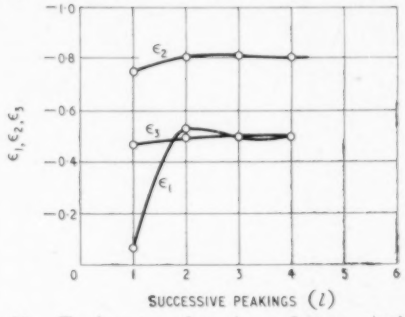


Fig. 15. *Tuning procedure in a 3-stage single-tuned amplifier* ($\alpha = 0{\cdot}5$). *Initially* $\epsilon_2 = 2$, $\epsilon_3 = -1$.

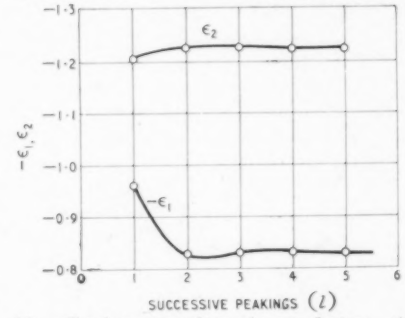


Fig. 16. *Tuning procedure in a 2-stage staggered amplifier* ($\alpha = 1$), $S(1^-, 1^+)$. *Initially* $\epsilon_2 = -4$.

frequency, while the second circuit tunes to another frequency; yet all the circuits trim to the frequencies higher than f_0. It may be pointed out that, when $\alpha = 1$, all the three circuits tune to $\epsilon = -1$ and the amplifier becomes unstable. In a four-stage amplifier, the first and the fourth circuits tune to one frequency, while the second and the third stages have their resonances at another frequency.

It is also found that, if the tuning procedure as described above is adopted, the circuits of a stagger-tuned amplifier do not tune to the theoretically computed frequencies. In a two-stage staggered amplifier both circuits tune to

frequencies higher than those applied, and the difference between them is somewhat greater than the difference between the two applied frequencies (see Fig. 16). The detuning from the desired frequencies depends on the magnitude of the regeneration coefficient α. It may be added that, in the case of a 3-stage staggered amplifier, the circuits will not tune to the desired frequencies and, furthermore, the resonant frequencies of the circuits will depend on the order of staggering.

It is possible, however, to tune the circuits to the desired frequencies if the following procedure is adopted: all the circuits of the amplifier, except the one which is to be trimmed, should be detuned far away from the centre frequency. Then, the particular circuit can be trimmed for maximum output at the indicator. It will tune (very nearly) to the applied frequency. With this procedure the signal at the output is small and, therefore, a receiver should be used as an output signal meter.

4. Conclusions

From the theory presented above it may be concluded that:

(i) For a variety of bandpass amplifiers it is possible to calculate the number of stages necessary to obtain a given bandwidth and gain. This can be done with the help of the formulae given in Section 2 and the curve in Fig. 2. The approach to amplifier design presented in that Section is valid only when regeneration is negligible.

(ii) In order to reduce regeneration, it is better to employ low dynamic impedances, rather than small mutual conductances for the valves, because α is proportional to the square of dynamic impedance, while the gain is directly proportional to it.

(iii) The effect of C_{ga} in single-tuned amplifiers is not as serious as it is in stagger-tuned amplifiers. A regeneration coefficient of 30% of the critical value may be permitted in single-tuned amplifiers without causing appreciable distortion of the shape of the response curves. For stagger-tuned amplifiers the regeneration coefficient α should be kept below 10% of the critical value, if response characteristics of good quality are to be obtained.

(iv) With regard to the problem of the best order of staggering, no definite conclusions have been reached by virtue of the theoretical investigation alone. Some experimental work has been done by the authors on three-stage and four-stage staggered amplifiers. On the strength of both the experimental results and the theory the following conclusions can be arrived at:

(a) It is inadvisable to put two high-Q circuits next to each other in a stagger-tuned amplifier;

(b) The first circuit of the amplifier should be tuned to a frequency below the centre frequency;

(c) If the regeneration coefficient is not too high, it is possible to compensate for the distortion of the response curves in stagger-tuned amplifiers by proper damping of anode circuits and appropriate choice of stagger frequencies. This technique may be of little value, however, in the manufacture of large quantities of amplifiers because the anode–grid capacitances and mutual conductances may vary from valve to valve. Thus, every single amplifier would have to undergo a lengthy process of compensation. It is advisable, therefore, to keep the regeneration coefficient low.

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APPENDIX 1

Analysis of Amplifier Composed of Double-plus-Single Tuned Groups.

The response curve of the type

$$K_m = \frac{K_{0m}}{(1 + x^{2m})^{\frac{1}{2}}} \qquad (1)$$

may be realized by staggering m single-tuned stages; the theory of stagger-tuned amplifiers is widely known and it will not be considered here. Another way of obtaining this type of response is as follows: a number of stages with double-tuned circuits of appropriate Q values and coupling coefficients k are cascaded. This arrangement is sufficient, if m is even; when m is odd, one single-tuned stage of appropriate bandwidth should also be included. This type of amplifier will now be analysed. It will be assumed that the inductances and the capacitances are equal for all circuits (including the single-tuned circuit).

The transfer impedance of a double-tuned inductively coupled circuit is

$$\mathbf{Z}_T = \frac{-j r_d Qk}{(Qk)^2 + 1 + 2jx_1 - x_1^2} = \frac{-j r_d p}{p^2 + 1 + 2jx_1 - x_1^2} \qquad (A1)$$

where

$$x_1 = Q_0 (f/f_0 - f_0/f) \approx \frac{Q_0}{f_0} 2\Delta f$$

and

$$Qk \approx Q_0 k = p$$

On the other hand it is known that the function

$$(1 + x^{2m})^{\frac{1}{2}} = \left[1 + (2\Delta f/B_m)^{2m}\right]^{\frac{1}{2}} \qquad (A2)$$

is the absolute value of

$$\prod_{h=1}^{h=\frac{m}{2}} \sin^2\left(\frac{2h-1}{2m}\right)\pi \left[1 + j\,\frac{2}{B_m \sin\left(\frac{2h-1}{2m}\right)\pi}\left\{\Delta f \pm \frac{B_m}{2}\cos\left(\frac{2h-1}{2m}\right)\pi\right\}\right] \quad \ldots \quad \ldots \quad \ldots \quad \text{(A3)}$$

if m is even; and it is the absolute value of

$$\left(1 + j\,\frac{2\Delta f}{B_m}\right) \times \prod_{h=1}^{h=\frac{m-1}{2}} \sin^2\left(\frac{2h-1}{2m}\right)\pi \times \left[1 + j\,\frac{2}{B_m \sin\left(\frac{2h-1}{2m}\right)\pi}\left\{\Delta f \pm \frac{B_m}{2}\cos\left(\frac{2h-1}{2m}\right)\pi\right\}\right] \ldots \quad \text{(A4)}$$

if m is odd.

Consider the general product:

$$\left[1 + j\,\frac{2}{B_m \sin\left(\frac{2h-1}{2m}\right)\pi}\left\{\Delta f + \frac{B_m}{2}\cos\left(\frac{2h-1}{2m}\right)\pi\right\}\right] \times \left[1 + j\,\frac{2}{B_m \sin\left(\frac{2h-1}{2m}\right)\pi}\left\{\Delta f - \frac{B_m}{2}\cos\left(\frac{2h-1}{2m}\right)\pi\right\}\right]$$

$$= 1 + \cot^2\left(\frac{2h-1}{2m}\right)\pi + j\,\frac{2 \times 2\Delta f}{B_m \sin\left(\frac{2h-1}{2m}\right)\pi} - \frac{4(\Delta f)^2}{B^2_m \sin^2\left(\frac{2h-1}{2m}\right)}\pi \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \text{(A5)}$$

By comparison of this product with the denominator of the transfer impedance (A1), it can be seen that in an amplifier with maximum-flatness response curve, every such product can be realized by employing one double-tuned stage, provided that the following conditions are fulfilled:

$$Q_h k_h = \cot\left(\frac{2h-1}{2m}\right)\pi$$

$$\text{and} \quad Q_h/f_0 = \frac{1}{B_m \sin\left(\frac{2h-1}{2m}\right)\pi} \quad \ldots \quad \ldots \quad \text{(A6)}$$

There are two cases:

(a) m—even

The response function (2) can be realized by using $\frac{1}{2}m$ double-tuned stages.

In order that the function (A2) in the denominator of the amplitude response may be obtained, the denominators of the constituent transfer impedances should be multiplied by the factors $\sin^2(\pi/2m)$, $\sin^2(3\pi/2m)$ . . . etc. Thus, the maximum gain of a group becomes

$$K_{0m} = g_m^{m/2} \times r_{d1} \times \cot\frac{\pi}{2m} \times \sin^2\frac{\pi}{2m} \times r_{d2} \times \cot\frac{3\pi}{2m} \times \sin^2\frac{3\pi}{2m} \times \ldots$$

where

$$r_{dh} = \frac{L}{R_h C} = \frac{Q_h}{2\pi f_0 C} = \frac{1}{2\pi C B_m \sin\left(\frac{2h-1}{2m}\right)\pi}$$

Therefore

$$K_{0m} = \left(\frac{g_m}{2\pi C B_m}\right)^{\frac{m}{2}} \times \prod_{h=1}^{\frac{m}{2}} \cos\left(\frac{2h-1}{2m}\right)\pi \quad \ldots \quad \text{(A7)}$$

But the product of cosines may be expressed as

$$s = \prod_{h=1}^{\frac{m}{2}} \cos\left(\frac{2h-1}{2m}\right)\pi = \frac{1}{2^{(m-1)/2}} \quad \ldots \quad \ldots \quad \text{(A8)}$$

Thus, finally

$$K_{0m} = \left(\frac{g_m}{2\pi C B_m}\right)^{\frac{m}{2}} \times 2^{-(m-1)/2} = \left(\frac{g_m}{2\pi C B_m}\right)^{\frac{m}{2}} \times s \quad \ldots \quad \ldots \quad \ldots \quad \text{(A9)}$$

The number of valves entering into a group is $t = \frac{1}{2}m$.

(b) m—odd

In this case the response function (1) may be realized by the use, in addition, of one single-tuned stage of bandwidth $B = B_m$. The total number of valves in the groups is $t = \frac{1}{2}(m+1)$ and the number of double-tuned coupled circuits is $\frac{1}{2}(m-1)$.

Maximum gain of a group is

$$K_{0m} = (g_m)^{(m+1)/2} \times r_{d1} \times r_{d2} \times \cot\frac{\pi}{2m} \times \sin^2\frac{\pi}{2m} \times r_{d3} \times \cot\frac{3\pi}{2m} \times \sin^2\frac{3\pi}{2m} \ldots$$

$$= \left(\frac{g_m}{2\pi C B_m}\right)^{\frac{m+1}{2}} \times \prod_{h=1}^{\frac{m-1}{2}} \cos\left(\frac{2h-1}{2m}\right)\pi$$

The product of cosines may be expressed as

$$s = \prod_{h=1}^{\frac{m-1}{2}} \cos\left(\frac{2h-1}{2m}\right)\pi = \frac{\sqrt{m}}{2^{(m-1)/2}} \quad \ldots \quad \text{(A11)}$$

And, finally $K_{0m} = \left(\frac{g_m}{2\pi C B_m}\right)^{\frac{m+1}{2}} \times \frac{\sqrt{m}}{2^{(m-1)/2}}$

$$= \left(\frac{g_m}{2\pi C B_m}\right)^{\frac{m+1}{2}} \times s \quad \ldots \quad \text{(A12)}$$

APPENDIX 2

Analysis of Stability of Bandpass Amplifiers

(a) *Single-Tuned Amplifier*

After the substitution of anode admittances (33) into the determinant of the network matrix (31), one obtains

$$\mathbf{D}_n = \frac{\mathbf{D}_{an}}{r_d{}^n} =$$

$$\frac{1}{r_d{}^n} \times \begin{vmatrix} 1+jx & -jb & 0 & 0 & \ldots & 0 & 0 \\ A & 1+jx & -jb & 0 & \ldots & 0 & 0 \\ 0 & A & 1+jx & -jb & \ldots & 0 & 0 \\ \ldots & \ldots & \ldots & \ldots & \ldots & \ldots & \ldots \\ 0 & 0 & 0 & \ldots & A & 1+jx & -jb \\ 0 & 0 & 0 & \ldots & 0 & A & 1+jx \end{vmatrix} \quad \text{(A13)}$$

where $A = g_m r_d$; $b = \omega C_{ga} r_d \approx \omega_0 C_{ga} r_d$

The determinant $\mathbf{D}_{an}$ can also be written in the form:—

$$\mathbf{D}_{an} = \begin{vmatrix} 1+jx & -j\alpha & 0 & 0 & . & 0 & 0 \\ 1 & 1+jx & -j\alpha & 0 & . & 0 & 0 \\ 0 & 1 & 1+jx & -j\alpha & . & 0 & 0 \\ . & . & . & . & . & . & . \\ 0 & 0 & 0 & . & 1 & 1+jx & -j\alpha \\ 0 & 0 & 0 & . & 0 & 1 & 1+jx \end{vmatrix} \quad \ldots \quad .. \quad .. \quad \text{(A14)}$$

where $\alpha = Ab = \omega_0 C_{ga} g_m r_d^2$

The determinant $\mathbf{D}_{an}$ is effectively a complex number and may be written as

$$\mathbf{D}_{an}(\alpha, x) = a_n[C_1(\alpha, x) + jC_2(\alpha, x)]$$

where a_n is some factor which cannot be equal to zero.

The amplifier will become unstable when

$$\mathbf{D}_{an}(\alpha, x) = 0$$

or

$$C_1(\alpha, x) + j\,C_2(\alpha, x) = 0 \qquad .. \qquad .. \qquad .. \quad \text{(A15)}$$

Thus there are two simultaneous equations (A15). whose solutions will give the frequency of oscillations (or x_n) and α_n.

If $x_n = x_n(\alpha)$ is a root of one of the equations (A15), then its substitution into $\mathbf{D}_{an}(\alpha, x)$ will result in a single equation in α; i.e.,

$$\mathbf{D}_{an}(\alpha, x_n) = a_n \begin{Bmatrix} C_1(\alpha, x_n) \\ +\,j\,C_2(\alpha, x_n) \end{Bmatrix} = 0 \qquad .. \qquad .. \quad \text{(A16)}$$

Thus any x_n which reduces $\mathbf{D}_{an}$ to a single equation can be considered as a solution of one of the equations (A15). It turns out that, for any $n \geqslant 2$, $x_n = -1$ possesses this property and hence may be regarded as a solution. This may be shown as follows:—

Let

$$1 - j = \sqrt{2}\, e^{-j\pi/4}$$

$$-j\alpha = \alpha e^{-j\pi/2}$$

The determinant $\mathbf{D}_{an}\,(x = -1)$ may then be expressed as

$$\mathbf{D}_{an}(x = -1) = \begin{vmatrix} \sqrt{2}\,e^{-j\pi/4} & \alpha e^{-j\pi/2} & 0 & 0 & . & 0 & 0 \\ 1 & \sqrt{2}\,e^{-j\pi/4} & \alpha e^{-j\pi/2} & 0 & . & 0 & 0 \\ 0 & 1 & \sqrt{2}\,e^{-j\pi/4} & \alpha e^{-j\pi/2} & . & 0 & 0 \\ . & . & . & . & . & . & . \\ 0 & 0 & 0 & . & 1 & \sqrt{2}\,e^{-j\pi/4} & \alpha e^{-j\pi/2} \\ 0 & 0 & 0 & . & 0 & 1 & \sqrt{2}\,e^{-j\pi/4} \end{vmatrix} \quad \text{(A17)}$$

Now the procedure is as follows:—

(1) multiply the 1st column by $e^{+j\pi/4}$
(2) multiply the 2nd row by $e^{-j\pi/4}$
(3) multiply the 2nd column by $e^{+j\pi/2}$
(4) multiply the 3rd row by $e^{-j\pi/2}$
(5) multiply the 3rd column by $e^{+j1\pi}$
(6) multiply the 4th row by $e^{-j1\pi}$

The process is continued until, finally, one obtains

$$\mathbf{D}_{an}(x = -1) = e^{-j\frac{\pi}{4}n} \times \begin{vmatrix} \sqrt{2} & \alpha & 0 & 0 & . & 0 & 0 \\ 1 & \sqrt{2} & \alpha & 0 & . & 0 & 0 \\ 0 & 1 & \sqrt{2} & \alpha & . & 0 & 0 \\ . & . & . & . & . & . & . \\ 0 & 0 & 0 & . & 1 & \sqrt{2} & \alpha \\ 0 & 0 & 0 & . & 0 & 0 & \sqrt{2} \end{vmatrix}$$

$$= D_{bn} e^{-j\frac{\pi}{4}n} \quad .. \quad \text{(A18)}$$

D_{bn} is a real quantity and a function of α only; when equated to zero it will yield the value of α necessary to cause oscillations. Thus it can be assumed that $x_n = -1$ is, indeed, one of the roots of $\mathbf{D}_{an} = 0$; furthermore (as previously explained) this is a physically acceptable solution.

D_{bn} may be regarded as a difference equation, since

$$D_{bn} = \sqrt{2}\, D_{b(n-1)} - \alpha D_{b(n-2)}$$

or

$$D_{bn} - \sqrt{2}\, D_{b(n-1)} + \alpha D_{b(n-2)} = 0 \qquad .. \qquad .. \quad \text{(A19)}$$

This, when solved by standard methods, yields

$$D_{bn} = \frac{1}{2\sqrt{\frac{1}{2} - \alpha}} \left\{ \left[\frac{1}{\sqrt{2}} + \sqrt{\tfrac{1}{2} - \alpha} \right]^{n+1} - \left[\frac{1}{\sqrt{2}} - \sqrt{\tfrac{1}{2} - \alpha} \right]^{n+1} \right\} \quad .. \qquad .. \quad \text{(A20)}$$

which, when expanded by the binomial theorem, gives the polynomial

$$D_{bn} = \left\{ \binom{n+1}{2} \left(\frac{1}{2}\right)^{\frac{n}{2}} + \binom{n+1}{3} \times \left(\frac{1}{2} - \alpha\right) \left(\frac{1}{2}\right)^{\frac{n-2}{2}} + \binom{n+1}{5} \left(\frac{1}{2} - \alpha\right)^2 \times \left(\frac{1}{2}\right)^{\frac{n-4}{2}} + \ldots \right\} \quad .. \quad \text{(A21)}$$

The polynomial is ended with the term

$$\binom{n+1}{n+1} \times \left(\frac{1}{2} - \alpha\right)^{\frac{n}{2}} \times \left(\frac{1}{2}\right)^{\frac{n-n}{2}}$$

if n is even, and

$$\binom{n+1}{n} \times \left(\frac{1}{2} - \alpha\right)^{\frac{n-1}{2}} \times \left(\frac{1}{2}\right)^{\frac{n-(n-1)}{2}}$$

if n is odd.

The values of critical regeneration coefficients α_n can be found by solving the equation $D_{bn} = 0$. Only the lowest positive root of this equation should be considered.

It can be verified (from A21) that, when $n \to \infty$, $\alpha_n \to \frac{1}{2}$.

The gain of the amplifier is given by

$$\mathbf{K}_n = \frac{(-1)^n \prod^{n} \mathbf{g}_{mk}}{\mathbf{D}_n}$$

If it is assumed that all $\mathbf{g}_{mk} = \mathbf{g}_m$, and $\mathbf{D}_n$ is expressed in terms of x and α, then the amplitude response function is given by

$$K_n(x) = \frac{{g_m}^n}{D_n(\alpha, x)} \qquad .. \qquad .. \qquad .. \qquad .. \quad \text{(A22)}$$

where $D_n(\alpha, x)$ is the modulus of $\mathbf{D}_n(\alpha, x)$. The successive five response functions are given in Table IA.

(b) *Staggered Pair*

The two anode admittances may be expressed by

$$\frac{1}{\mathbf{Z}_{11}} = \frac{1 + j(\sqrt{2}x + 1)}{r_d} \qquad \text{(A23)}$$

$$\frac{1}{\mathbf{Z}_{22}} = \frac{1 + j(\sqrt{2}x - 1)}{r_d}$$

where $x = \sqrt{2}\,\Delta f/B'$

B' being the bandwidth of a single circuit.

The determinant of the network is

$$\mathbf{D}_2 = \frac{1}{r_d^2}[(2 - 2x^2 + j(2\sqrt{2}x + \alpha)] \qquad \text{(A24)}$$

and it is independent of the order of staggering.

The conditions of oscillation are

$$2 - 2x^2 = 0$$

$$2\sqrt{2}x + \alpha = 0 \qquad \text{(A25)}$$

$$K_{11}(\alpha, x) = \frac{\frac{1}{4}(g_m r_d)^4}{\sqrt{\left(1 + x^4 - 4x^2 - \frac{3}{2}\sqrt{2}\alpha x - \frac{\alpha^2}{4}\right)^2 + \left[(1 - x^2)\left(2\sqrt{2}x + \frac{3}{2}\alpha\right)\right]^2}} \qquad \text{(A29)}$$

while for $S_3(1^-, 1^+, 1^+, 1^-)$ it is

$$K_{31}(\alpha, x) = \frac{\frac{1}{4}(g_m r_d)^4}{\sqrt{\left[1 + x^4 - 4x^2 - \frac{\alpha}{4}(6\sqrt{2}x + \alpha + 2)\right]^2 + \left[2\sqrt{2}x(1 - x^2) + \frac{\alpha}{4}(4 - 6x^2 - 2\sqrt{2}x)\right]^2}} \qquad \text{(A30)}$$

which yield $x_2 = -1 \quad ; \quad \alpha_2 = 2\sqrt{2}$

The amplitude response curve is

$$K_2(\alpha, x) = \frac{\frac{1}{2}(g_m r_d)^2}{\sqrt{1 + x^4 + \sqrt{2}\,\alpha x + \alpha^2/4}} \qquad \text{(A26)}$$

(c) *m Staggered Pairs*

The admittances of the anode circuits are

$$\frac{1}{\mathbf{Z}_{kk}} = \frac{1 + j(\sqrt{2}\,x + 1)}{r_d} \qquad \text{(A27)}$$

$$\frac{1}{\mathbf{Z}_{jj}} = \frac{1 + j(\sqrt{2}\,x - 1)}{r_d}$$

where $x = \sqrt{2}\,\Delta f/B'$

Consider, first, two cascaded staggered pairs. For this case the determinant of the network is

$$\mathbf{D}_4 = \frac{1}{\mathbf{Z}_{11}\mathbf{Z}_{22}\mathbf{Z}_{33}\mathbf{Z}_{44}} + jg_m\omega C_{ga}\left(\frac{1}{\mathbf{Z}_{11}\mathbf{Z}_{22}} + \frac{1}{\mathbf{Z}_{11}\mathbf{Z}_{44}} + \frac{1}{\mathbf{Z}_{33}\mathbf{Z}_{44}}\right) - (g_m\omega C_{ga})^2 \qquad \text{(A28)}$$

Again, the conditions of oscillation of two staggered pairs may be found by solving the equation $\mathbf{D}_4 = 0$.

For the case $S_1(1^-, 1^+, 1^-, 1^+) \equiv S_2(1^+, 1^-, 1^+, 1^-)$ the amplitude response function is found to be

Now m 'interlaced' staggered pairs will be considered and the conditions of oscillation will be found. The 'interlaced' staggering order is represented by

$$S_a(1^+, 1^-, 1^+, 1^- \ldots) = S_b(1^-, 1^+, 1^-, 1^+ \ldots)$$

If the anode admittances are expressed in the form shown in (A27), then the determinant of the network may be written

$$D_{2m} = \frac{1}{r_d^{2m}} \begin{vmatrix} 1 + j(\sqrt{2}x - 1) & -j\alpha & 0 & 0 & \cdot & 0 & 0 \\ 1 & 1 + j(\sqrt{2}x + 1) & -j\alpha & 0 & \cdot & 0 & 0 \\ 0 & 1 & 1 + j(\sqrt{2}x - 1) & -j\alpha & \cdot & 0 & \cdot \\ \cdot & \cdot & \cdot & \cdot & \cdot & \cdot & \cdot \\ \cdot & \cdot & \cdot & \cdot & \cdot & \cdot & \cdot \\ 0 & 0 & \cdot & \cdot & 1 & 1 + j(\sqrt{2}x - 1) & -j \\ 0 & 0 & \cdot & \cdot & 0 & 1 & 1 + j(\sqrt{2}x + 1) \end{vmatrix} \qquad \text{(A31)}$$

TABLE 1A

n	Amplitude Response $K_n(\alpha, x)$
1	$\dfrac{g_m r_d}{\sqrt{1 + x^2}}$
2	$\dfrac{(g_m r_d)^2}{\sqrt{(1 - x^2)^2 + (2x + \alpha)^2}}$
3	$\dfrac{(g_m r_d)^3}{\sqrt{(1 + x^2)\,[(1 - x^2)^2 + 4(x + \alpha)^2]}}$
4	$\dfrac{(g_m r_d)^4}{\sqrt{[x^4 - 6x^4 - \alpha(\alpha + 6x) + 1]^2 + [4x - 4x^3 - 3\alpha(x^2 - 1)]^2}}$
5	$\dfrac{(g_m r_d)^5}{\sqrt{(1 + x^2)\,\{[x^4 - 6x^2 - \alpha(8x + 3\alpha) + 1]^2 + [x - x^3 + \alpha(1 - x^2)]^2 \times 16\}}}$

As in Appendix 2(a), the substitution of $x_{2m} = -1$ into the determinant (A31), results in a single equation, which, when solved, gives the second condition of oscillation $\alpha = \alpha_{2m}$. This is proved as follows:

Let

$$1 - j(\sqrt{2} + 1) = a_1 e^{-j\theta_1},$$
$$1 + j(-\sqrt{2} + 1) = a_2 e^{-j\theta_2} \quad .. \quad .. \quad (A32)$$

and $\quad -j\alpha = \alpha e^{-j\pi/2}$

It follows from Equ. (A32) that

$$a_1 a_2 = 2\sqrt{2}$$
$$\theta_1 + \theta_2 = \pi/2$$

Now $x = -1$ is substituted into the determinant (A31) and a number of algebraic operations are performed, during which the properties $\theta_1 + \theta_2 = \pi/2$ and $a_1 a_2 = 2\sqrt{2}$ are made use of. This leads to an equation in $\alpha/\sqrt{2}$, similar to Equ. (37) and thus, the critical regeneration coefficients for this case are given by

$$\alpha_{2m} = \sqrt{2}\alpha_n \quad .. \quad .. \quad .. \quad .. \quad (A33)$$

where only even n should be considered.

(d) *Staggered Triple*

The three anode admittances which enter into a staggered triple may be represented as

$$\frac{1}{\mathbf{Z}_a} = \frac{1 + j(2x + \sqrt{3})}{2r_d} \qquad \text{high-}Q \text{ circuit}$$
$$\frac{1}{\mathbf{Z}_b} = \frac{1 + j(2x - \sqrt{3})}{2r_d} \qquad \text{high-}Q \text{ circuit} \quad .. \quad (A34)$$
$$\frac{1}{\mathbf{Z}_c} = \frac{1 + jx}{r_d} \qquad \text{low-}Q \text{ circuit}$$

where $x = 2\Delta f/B_3$

B_3 is the bandwidth of the low-Q circuit, which is also the overall bandwidth of the staggered triple under the condition of negligible regeneration.

r_d is the dynamic impedance of the low-Q circuit.

For the case $S_1(1^-, 0, 1^+) \equiv S_2(1^+, 0, 1^-)$ the determinant of the network becomes

$$\mathbf{D}_3 = \frac{1}{4r_d{}^3}[4 - 8x^2 - 8\alpha x + j(4\alpha - 8x - 4x^3)] \quad (A35)$$

Thus the conditions for oscillation are

$$4 - 8x^2 - 8\alpha x = 0$$
$$4\alpha - 8x - 4x^3 = 0 \quad .. \quad .. \quad .. \quad (A36)$$

which yield

$$x_1 = -\sqrt{\tfrac{1}{2}(1 + \sqrt{3})} = -1{\cdot}17$$
$$\alpha_3 = \frac{1}{2}\frac{\sqrt{6}}{\sqrt{1 + \sqrt{3}}} = 0{\cdot}743$$

The amplitude response function is given by

$$K_3(x, \alpha) = \frac{(g_m r_d)^3}{\sqrt{1 + x^6 + 6\alpha x^3 + 4\alpha^2 x^2 + \alpha^2}} \quad .. \quad (A37)$$

The second case is $S_3(0, 1^-, 1^+) \equiv S_4(1^+, 1^-, 0)$, for which the determinant of the network is in the form

$$\mathbf{D}_3 = \frac{1}{4r_d{}^3}[4 - 8x^2 - 8\alpha x + 2\sqrt{3}\alpha + j(8x - 4x^3 + 6\alpha)]$$
$$.. \quad .. \quad .. \quad (A38)$$

Here the oscillation conditions are given by

$$4 - 8x^2 - 8\alpha x + 2\sqrt{3}\alpha = 0 \quad .. \quad .. \quad (A39)$$
$$8x - 4x^3 + 6\alpha = 0$$

which yield

$$x_1 = -1{\cdot}15$$
$$\alpha_3 = 0{\cdot}518$$

The amplitude response function is

$$K_3(x, \alpha) = (g_m r_d)^3 \div$$
$$\sqrt{\left[(1 + \sqrt{3}\alpha + 3\alpha^2) + 2\alpha(1 - \sqrt{3}\alpha)x + \right.}$$
$$\left. 2\alpha(2\alpha - \sqrt{3})\,x^2 + 5\alpha x^3 + x^6\right] \quad .. \quad .. \quad (A40)$$

Finally, the last case, $S_5(1^-, 1^+, 0) \equiv S_6(0, 1^+, 1^-)$ has the determinant

$$\mathbf{D}_3 = \frac{1}{4r_d{}^3}[4 - 2\alpha\sqrt{3} - 8\alpha x - 8x^2 + j(8x - 4x^3 + 6\alpha)]$$
$$.. \quad .. \quad .. \quad .. \quad (A41)$$

which is equal to zero, when

$$x_1 = -0{\cdot}916$$
$$\alpha_3 = 0{\cdot}702$$

The response function becomes

$$K_3(x, \alpha) = (g_m r_d)^3 \div$$
$$\sqrt{\left[(1 - \sqrt{3}\alpha + 3\alpha^2) + 2\alpha(1 + \sqrt{3}\alpha)x + \right.}$$
$$\left. 2\alpha(2\alpha + \sqrt{3})\,x^2 + 5\alpha x^3 + x^6\right] \quad .. \quad .. \quad (A42)$$

SINGLE-STAGE PHASE-SHIFT OSCILLATOR

Method of Design

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SUMMARY.—It is pointed out that there is no special advantage in making half the components of the phase-shift network of the same value, as is frequently done. A different choice of relative values leads to a network requiring less gain from the valve and enables formulae to be obtained which apply to a network having any number of sections. Some general conclusions as to the suitability of various networks for different applications are drawn from these formulae, and values of frequency and network input impedance are tabulated.

Introduction

SINCE the original description in 1923,[1] a number of papers dealing with the behaviour of the phase-shift oscillator have appeared.[2] The majority of these describe circuits in which feedback takes place through a network of the form shown in Fig. 1 (a) and (b). At one particular frequency such a network will produce a phase shift of 180° and it is approximately at this frequency that oscillation takes place. At (a) and (b) of Fig. 1 are shown oscillators made up of three basic sections of the type of (c) and (d). There does not appear to be any fundamental reason why an oscillator should not function with any number of cells greater than three, but it is common practice to use not more than four. Formulae have been given for these cases;[3,4] the work involved in obtaining these is, however, somewhat laborious and increases rapidly as the number of sections increases. The usual design procedure is to make all the resistors of the same value and the capacitors similarly all of the same reactance, when simpler expressions for the frequency of oscillation and gain may be obtained.

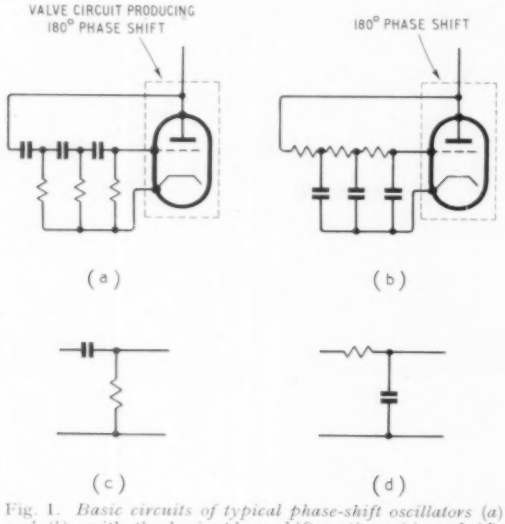

Fig. 1. *Basic circuits of typical phase-shift oscillators (a) and (b), with the basic phase-shift sections (c) and (d).*

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If ganged components are being used for frequency variation it is obviously desirable to have either all the reactances or all the resistances identical, but it is not essential to have both. Fig. 2 shows a three-cell network in which one of the components is made N times the value of the others. A paper by Vaughan[5] considers the effect on the gain required for oscillation of varying the value of N; curves given indicate that the minimum value is not obtained when the components are equal. It is therefore interesting to find the value of N which does require least gain. The expression given by Vaughan for the ratio of input voltage to output voltage for the network of Fig. 2 (a) is

$$V_L/V_a = -[14 + 12N + 3/N]$$

and for Fig. 2 (b)

$$V_L/V_a = -[14 + 3N + 12/N]$$

Differentiating the first equation with respect to N and equating to zero yields the value $N = \frac{1}{2}$, while the second equation yields the value $N = 2$. Thus, in either case the best result is obtained if the first series impedance is of half the value of the succeeding ones. A consideration of the 4-cell circuit shows that an improvement is obtained by halving the first series impedance in this case also.

It therefore seems reasonable to develop a method of design in which the first series impedance has half the value of the others. By so doing hyperbolic formulae can be obtained which are readily applicable to networks containing any number of cells.

Development of Formulae

Consider a three-section RC network with the first series resistance of value $R/2$. Since the network is assumed working into an open circuit, the introduction of a further series resistance $R/2$ at the end of the network as shown in Fig. 3(a) will not in any way affect the properties of the circuit. We may now, however, split up the ladder into a chain of three T-links, as shown in Fig. 3(b). The same procedure could be used for a network containing any number of sections n, in which case n T-links would have been obtained.

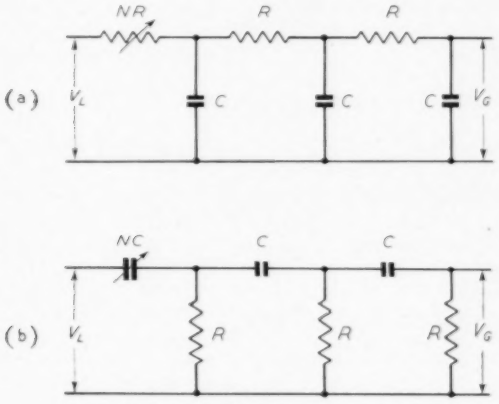

Fig. 2. *Three-cell phase-shift networks having one component N times the value of the others.*

Now it is well known[6] that for such a chain of T-links output voltage on open circuit is related to input voltage by the expression

$$V_G = V_L/\cosh n\theta$$

In this case $\cosh \theta = 1 + \frac{1}{2}R.j\omega C = 1 + j\omega CR/2$. θ will, in general, be complex and so we may write $\theta = \alpha + j\beta$. The condition of interest is the one in which V_L and V_G are in antiphase; i.e., $\cosh n\theta$ has no imaginary component. This leads (see Appendix 1) to the relationship $\beta = \pi/n$. Expanding $\cosh \theta$ we obtain

$$\cosh \theta = \cosh (\alpha + j\beta) = \cosh \alpha \cos \beta + j \sinh \alpha \sin \beta = 1 + j\omega CR/2$$

and equating real and imaginary parts yields the result

$$\omega CR/2 = \sinh \alpha \sin \beta = \tan \beta \sin \beta = \tan (\pi/n) \sin (\pi/n)$$

Whence

$$\omega = [2 \sin (\pi/n) \tan (\pi/n)]/CR$$

or $f = \sin (\pi/n) \tan (\pi/n)/\pi CR$

The input/output voltage ratio is also obtained as

$$V_L/V_G = \cosh n\theta = \cosh \{n \cosh^{-1} [1/\cos (\pi/n)]\}$$

On applying the same reasoning to a ladder with series capacitors and shunt resistors, with the first capacitor of value $2C$ [Fig. 3(c) and 3(d)] it is found that the input/output voltage ratio is unchanged but the frequency is now given by

$$f = 1/[4\pi CR \sin (\pi/n) \tan (\pi/n)] \text{ or}$$

$$\omega = 1/[2 \sin (\pi/n) \tan (\pi/n) CR]$$

The expressions for ω in these two cases may alternatively be written

$$\omega = K/CR \text{ and}$$

$$\omega = 1/KCR \text{ where } K = 2 \sin (\pi/n) \tan (\pi/n)$$

Input Impedance

The above method also enables formulae to be obtained for the input impedance of the network. For an open-circuited chain of T-links, $Z_{IN} = Z_0/\tanh n\theta$ and at the frequency of oscillation this becomes $Z_{IN} = Z_0/\tanh n\alpha$. For the values of α which are obtained, $\tanh n\alpha$ is so close to unity that it may be removed from the equation. Then for the RC network

$$Z_{IN} = Z_0/\tanh \alpha n = Z_0 = \frac{R}{2}\sqrt{1 - 4j/K}$$

and for the CR network

$$Z_{IN} = Z_0/\tanh \alpha n = Z_0 = [RK/2j]\sqrt{1 + 4j/K}$$

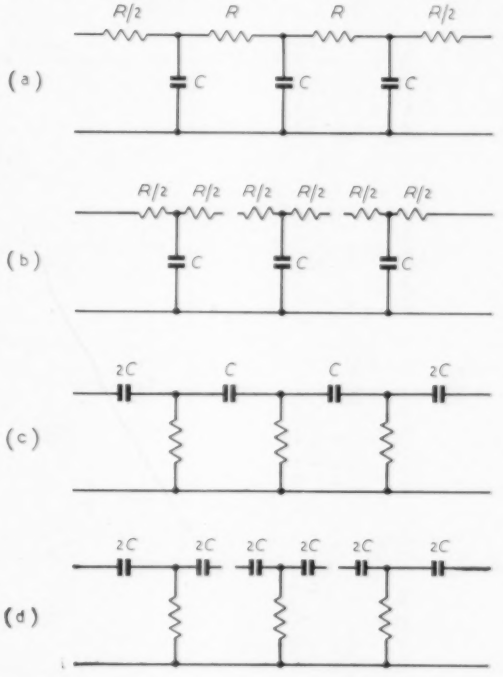

Fig. 3. *This diagram shows how three-stage T-section networks can be broken up into identical sections.*

Alternative Ladder Forms

Instead of halving the first series impedance, the last shunt impedance may be doubled, and an additional shunt impedance of value equal to the last one added on at the beginning of the network. Omission of this first impedance would not affect the formulae for frequency but would alter the input impedance. An RC network, as shown in Fig.4(a), now results and a CR network, as shown in Fig. 4(c). These may be split up in the way shown in (b) and (d), but this time a chain of π-links is obtained. The frequency and voltage ratio formulae are unchanged, but the impedance formulae now become

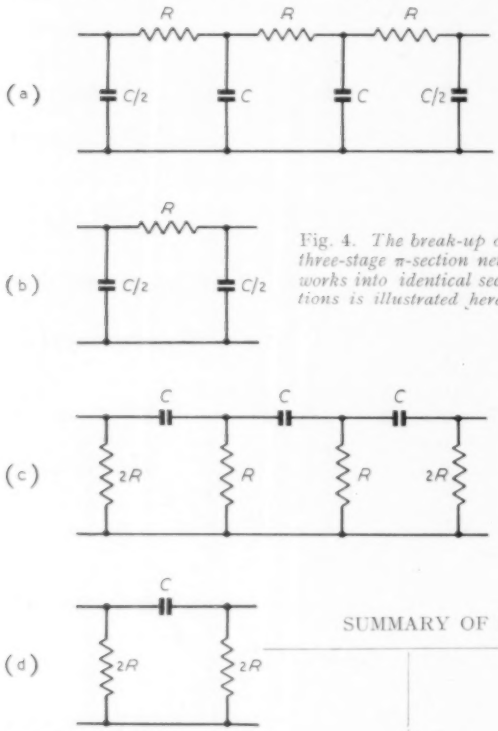

Fig. 4. *The break-up of three-stage π-section networks into identical sections is illustrated here.*

$$Z_{IS} = 2R/[jK\sqrt{1 - 4j/K}]$$

and $$Z_{IS} = 2R/\sqrt{1 + 4j/K}$$

All these results are summarized in Tables 1 and 2, where values are given for oscillators using up to 6 sections.

Practical Verification

To obtain a practical verification of these results, the circuit shown in Fig. 5 was set up. Figures obtained were as follows:

Number of sections	3	4	5	6
Frequency (measured)	108	227	373	545 c/s
Frequency (calculated)	106	225	374	550 c/s

It was found that the variation of input/output voltage ratio for the different networks caused a change in output voltage and variation of frequency. The grid lead was accordingly taken to a tapping on the last shunt resistor, the position of which was adjusted as the number of sections was varied so as to maintain output constant at a low value. Variation in h.t. voltage of 30% was found to alter the frequency by less than 1% in the 6-section case, though rather more with 3 sections.

To obtain an idea of the effects of stray capacitance and interelectrode capacitance, an extra 120 pF was connected from grid to earth; the change of frequency was very small, of the order of 1 c/s at 550 c/s. The addition of 1 kΩ in series with the first series impedance caused the frequency to fall from 545 to 541 c/s. The impedance of the cathode follower was assessed at 500 ohms and would, therefore, have only a very small effect on the frequency of oscillation.

These measurements indicated the extent of the most probable errors, and a satisfactory degree of agreement was obtained between theoretical and practical results, as shown by the figures above.

TABLE 1

SUMMARY OF FREQUENCY AND IMPEDANCE FORMULAE

		ω	Input impedance	
			First series impedance $= R/2$ or $2C$	Last shunt impedance $= 2R$ or $C/2$
Exact Formulae	RC oscillator	K/CR	$[R/2]\sqrt{1 - 4j/K}$	$2R/jK\sqrt{1 - 4j/K}$
	CR oscillator	$1/KCR$	$R/2.K/j.\sqrt{1 + 4j/K}$	$2R/\sqrt{1 + 4j/K}$
Approximate Formulae	RC oscillator	K_1^2/CR	R/K_1	R/K_1
	CR oscillator	$1/K_1^2CR$	K_1R	K_1R

$K = 2 \sin [\pi/n] \tan [\pi/n]$; $K_1 = \pi\sqrt{2}/n$

Approximate Formulae for Large Values of n

If n is large, approximate formulae may be obtained by writing $\sin [\pi/n] = \tan [\pi/n] = \pi/n$. The formulae for frequency then become

$$f = 1/[4\pi CR(\pi/n)\ (\pi/n)] = [n^2/4\pi^3]/CR$$
$$\text{or} \quad \omega = [n^2/2\pi^2]/CR$$

for the CR oscillator, and

$f = [\pi/n^2]/CR$ or $\omega = [2\pi^2/n^2]/CR$

for the RC oscillator. Approximate formulae for the impedance may similarly be obtained (see appendix 2) and are listed in Table 2. It will be seen that, provided n is greater than 4, the approximate formulae give reasonable results, the error when n is equal to 4 being about 13%.

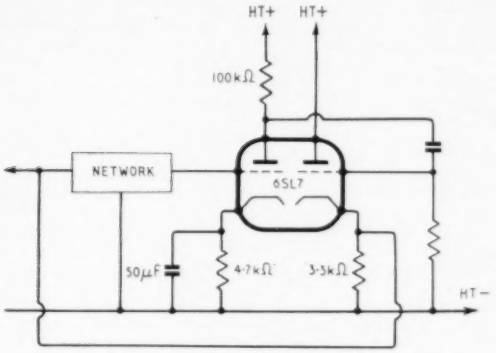

Fig. 5. *Experimental phase-shift oscillator circuit.*

Thus, in the first case, ω decreases with the number of sections and, in the second case, increases with the number of sections. If the design of an oscillator were being considered in which it was desired to obtain the highest possible frequency with a fixed minimum value of capacitance, the CR network with a large value of n would have advantages. If, on the other hand, n is restricted to the value 3 or 4, the RC network will give the greater frequency.

In the case of a very low-frequency oscillator it is usually desired to obtain the lowest possible frequency for a given total capacitance employed. In the RC network frequency is independent of the total capacitance (nC) employed for a given Z_{IN}, but this is not so for the CR network in which ω decreases with n. Taking the lowest possible value, n equals 3, shows that the CR network yields a lower frequency than the RC network, and a three-section CR network is therefore the best choice.

Acknowledgment is made to the Governors of the Municipal College, Portsmouth, who enabled this investigation to be carried out.

TABLE 2

VALUES OF IMPEDANCE AND FREQUENCY COEFFICIENTS FOR n from 1 to 6

n	k	k_1^2	Magnitude of Z/R					
			RC oscillator			CR oscillator		
			½ series input	½ shunt input	approx.	½ series input	½ shunt input	approx.
3	3	2·19	0·646	0·516	0·675	1·938	1·55	1·48
4	1·414	1·233	0·866	0·817	0·9	1·22	1·15	1·11
5	0·853	0·79	1·095	1·07	1·125	0·935	0·913	0·89
6	0·577	0·55	1·325	1·31	1·35	0·764	0·756	0·74

Conclusion

The foregoing work results in simple formulae for the impedance of the networks concerned and the frequency of oscillation. The relative simplicity of these formulae, whatever the value of n, enables an assessment to be made as to the suitability of different networks for different purposes.

Usually the oscillator must be designed with regard to various limiting factors. One of these will nearly always be the input impedance of the network, which must not fall below a specified value in order to prevent undue loading of the valve. Using the approximate formulae we may express the frequency of oscillation in terms of the impedance, namely: $\omega = \sqrt{2}\pi/(nCZ_{IN})$ for the RC section and $\omega = n/(\sqrt{2}\pi CZ_{IN})$ for the CR.

LIST OF SYMBOLS

V_L = input voltage to network
V_θ = output voltage from network
n = number of sections in network
θ = transfer constant per section $= \alpha + j\beta$
$\cosh\theta = 1 + Z_1/2Z_2$
Z_1 = total series impedance per cell
Z_2 = total shunt impedance per cell
Z_0 = iterative impedance of network
Z_{IN} = input impedance of network
$K = 2\sin[\pi/n]\tan[\pi/n]$; $K_1 = \pi\sqrt{2}/n$
R = resistance/cell of network
C = capacitance/cell of network

APPENDIX 1

Derivation of Formulae for Oscillation Frequency and Gain Required

$V_\theta = V_L / \cosh n\theta$
$= V_L / [\cosh \alpha n \cos \beta n + j \sinh \alpha n \sin \beta n]$

If the two voltages are to be in antiphase, the imaginary

term must disappear. Hence $\sinh \alpha n \sin \beta n = 0$ and since α cannot be zero, $\sin \beta n = 0$. Hence $\beta n = \pi$ (though other solutions are mathematically possible, this is the one which corresponds with the physical reality) and $\beta = \pi/n$. But $\cosh \theta = 1 + Z_1/2Z_2$. Hence for the RC oscillator network of Fig. 3(a): $\cosh \theta = 1 + j\omega CR/2$. Expanding $\cosh \theta$ and equating reals and imaginaries yields $\cosh \alpha \cos \beta = 1$ and $\sinh \alpha \sin \beta = \omega CR/2$. From the first equation $\cosh \alpha = 1/\cos \beta$ whence

$$\sinh \alpha = \sqrt{\cosh^2 \alpha - 1} = \sqrt{1/\cos^2 \beta - 1}$$

$$= \sqrt{\frac{1 - \cos^2 \beta}{\cos^2 \beta}} = \tan \beta$$

Thus $\tan \beta \sin \beta = \omega CR/2$ whence $\omega = 2 \tan \beta \sin \beta / CR$

$$\therefore \omega = 2 \tan [\pi/n] \sin [\pi/n] / CR$$

and if $K = 2 \tan [\pi/n] \sin [\pi/n]$ then $\omega = K/CR$

For the CR oscillator network of Fig. 3(c) the same reasoning applies with the exception that now $\cosh \theta = 1 + 1/(2j\omega CR)$.

Hence $\tan \beta \sin \beta = \frac{1}{2}\omega CR$ and

$$\omega = 1/[2 \tan (\pi/n) \sin (\pi/n) CR] = 1/(KCR)$$

The voltage ratio in either case is given by

$$V_L/V_G = \cosh n\theta = -\cosh n\alpha$$

$$= -\cosh [n \cosh^{-1} (1/\cos \pi/n)]$$

APPENDIX 2

Derivation of Impedance Formulae

It is well known that for the T-network of Fig. 6(a) iterative impedance is given by the formula

$$Z_0 = \sqrt{Z_1 Z_2 + Z_1^2/4} = \sqrt{Z_1 Z_2 (1 + Z_1/4Z_2)}$$

and for the π network of Fig. 6(b)

$$Z_0 = \sqrt{Z_1 Z_2/(1 + Z_1/4Z_2)}$$

Thus the iterative impedance of the RC ladder arranged in T sections as in Fig. 3(b) is obtained by putting $Z_1 = R$ and $Z_2 = 1/j\omega C$ then $Z_0 = \sqrt{R/j\omega C + R^2/4}$ Taking R outside the square-root sine and substituting $\omega CR = K$ gives

$$Z_0 = R\sqrt{1/jK + 1/4} = [R/2]\sqrt{1 - 4j/K}$$

But for an open-circuited chain of networks, input impedance is given by $Z_{IN} = Z_0/\tanh m\theta$; now $\tanh m\theta = \tanh (n\alpha + jn\beta)$ and since at the oscillation frequency $n\beta = \pi$, this reduces to $\tanh n\alpha$. α is, however, so large that $\tanh n\alpha$ is equal to unity with better than $\frac{1}{2}\%$ approximation. For the network of Fig. 3(a) we thus obtain

$$Z_{IN} = Z_0 = [R/2]\sqrt{1 - 4j/K}$$

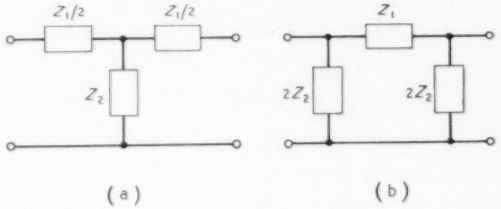

Fig. 6. *Networks employed in deriving the characteristics of the sections.*

Formulae for the other cases may be similarly obtained. When the number of sections is very large

$$K = 2 \tan [\pi/n] \sin [\pi/n] \rightarrow 2\pi^2/n^2$$

Thus in the limit $Z_{IN} = [R/2]\sqrt{1 - 2jn^2/\pi^2}$ which in magnitude tends to $Z_{IN} = [R/2]n\sqrt{2}/\pi = K_1 R/2$ where $K_1 = n\sqrt{2}/\pi$

REFERENCES

[1] U.S. Patent No. 1,442,871. H. W. Nicholls, 16th January, 1923.

[2] Most of these after 1941; see, for example, Ginzton and Hollingsworth, "Phase Shift Oscillators", *Proc. Inst. Radio Engrs*, February 1941.

[3] Parton and Roorda, correspondence, *Electronic Engineering*, January 1950.

[4] Hinton, "Design of R-C Oscillator Phase-Shift Network", *Electronic Engineering*, January 1950.

[5] W. C. Vaughan, "Phase Shift Oscillator", *Wireless Engineer*, December 1949.

[6] Mallett, "Telegraphy and Telephony", or other standard textbook.

BRIT.I.R.E. CONVENTION

The 1954 Convention of the British Institution of Radio Engineers is being devoted to the application of electronics to industrial control, processes and computation. It is to be held in Christ Church, University of Oxford, from 8th July to 12th July.

The six sessions will cover industrial applications of electronic computors, industrial applications of X-rays and ultrasonics for testing, Radioactive devices for testing, Electronic sensing devices, Electronic actuators and Electronic aids to production.

MEETINGS

I.E.E.

7th April. "A Versatile Transistor Circuit", by E. H. Cooke-Yarborough, M.A. "The Measurement of the Small-Signal Characteristics of Transistors", by E. H. Cooke-Yarborough, M.A., C. D. Florida and J. H. Stephen, Ph.D. "A Bridge for Measuring the A.C. Parameters of Type 'A' Transistors", by A. R. Boothroyd, Ph.D., and L. K. Datta, M.Sc. "The Transistor as a Regenerative Amplifier with some Application to Computing Circuits", by G. B. B. Chaplin, M.Sc.

12th April. "Safety Measures for Radio and Television Equipment", discussion to be opened by E. P. Wethey.

23rd April. "A Method for the Plotting of Magnetic Fields due to Currents", discussion to be opened by F. T. Chapman, C.B.E., D.Sc.(Eng.), and D. G. Sandeman, B.Sc.(Eng.).

26th April. "The Experimental Synthesis of Speech", by W. Lawrence.

27th April. "Auto-Self-Excited Transductors and Push-Pull Circuit Theory", by A. G. Milnes, M.Sc., and T. S. Law, B.Sc. "Composite Cores for Instrument Transductors", by E. H. Frost-Smith, B.A., Ph.D., and A. E. De Barr, B.Sc.

29th April. "The Physics of the Ionosphere", Forty-fifth Kelvin Lecture by J. A. Ratcliffe, O.B.E., M.A., F.R.S.

5th May. "The Reflection and Absorption of Radio Waves in the Ionosphere", by W. R. Piggott, B.Sc. "Some Notes on the Absorption of Radio Waves Reflected from the Ionosphere at Oblique Incidence", by W. J. G. Beynon, Ph.D., D.Sc.

These lectures will be held at the Institution of Electrical Engineers, Savoy Place, Victoria Embankment, London, W.C.2, and will commence at 5.30.

Brit.I.R.E.

21st April. "Crystal Valves in Radio and Electronics", by B. R. A. Bettridge, to be held at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1, commencing at 6.30.

CORRESPONDENCE

Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

Constant-H.T. Current Amplifier

SIR,—I was interested to read your Editorial in the February 1954 issue of *Wireless Engineer*.

The amplifier circuit you describe, consisting of an amplifier stage followed by a cathode-follower stage, with direct coupling and having load resistances substantially equal in order to present a constant load to the h.t. supply, has been of considerable service, particularly in video amplifiers for television transmitters where modulation is performed at high level.

This application, still in use in the Alexandra Palace transmitter, is described on p. 774 of the classic paper on "The Marconi-E.M.I. Television System", by Blumlein, Browne, Davis and Green, *J. Instn elect. Engrs*, Vol. 83, p. 758, December 1938. See also British Patent 462,536.

The arrangement was again used in the Sutton Coldfield transmitter, and a fairly full description and analysis is given in Section 7.3 of the paper by Nind and Leyton in *Proc. Instn elect. Engrs*, Vol. 98, Part III, p. 416, November 1951.

The conception of the circuit was largely due to the non-existence of pentode or tetrode valves of adequate power for these applications, and it will be noted that the arrangement is particularly well adapted to the use of cascaded units, since the large Miller input capacitance of the triode, to which you refer, is looked after by the low-impedance drive from the cathode-follower output stage of the preceding unit.

Another point helping satisfactory video response with a triode amplifier stage is that the impedance at the anode of V_1 is considerably less than R_3 since the r_a of V_1 is in parallel.

ERIC L. C. WHITE.

E.M.I. Research Laboratories, Ltd.,
Hayes, Middlesex.
19th February 1954.

Transmission-Line Matching System

SIR,—The paper by R. E. Collin and J. Brown in your February issue gives a complete theoretical treatment of two quarter-wave transformers used to match two transmission lines of different characteristic impedances.

A good approximate solution which the writer used exists for this problem and for the problem of any number of quarter-wave transformers. It may be worth while stating in this connection, especially as it gives some insight into the physical processes involved. Only a brief statement is possible here.

The method assumes the reflections at each discontinuity (change of Z_0) to be small. It is based on a method of designing phase-shifters due to G. J. Halford.[1] The problem there is to transform from an air-filled waveguide to one partially filled with dielectric and therefore of lower Z_0.

Denote the total reflection from the system by r. The transition consists of a number of quarter-wave sections having intermediate values of Z_0. The reflection coefficients at each discontinuity are in the ratio of the binomial coefficients if an optimum match at the design frequency is required, but different if two or more discrete zeros of r at different frequencies are desired. Thus with two transformers arranged binomially the individual reflection coefficients are as 1 : 2 : 1 to give a flat zero in r at a frequency f, say; but they become 1 : 2 cos α : 1 if r is to be zero at frequencies $f(\pi \pm \alpha)/\pi$ (and incidentally not zero at f). If the characteristic impedances are Z_{in}, Z_1, Z_2, and R (to use Collin and Brown's notation), then $Z_1/Z_{in} : Z_2/Z_1 : R/Z_2$ as 1 : 2 : 1 or as 1 : 2 cos α : 1 according to the bandwidth required. Halford quotes results for up to four transformers, and the writer has obtained similar results. A feature of this method is that the more transformers are used between two given characteristic impedances, the less are the individual reflections and hence the more accurate the theory becomes (but the more critical the tolerances).

Collin and Brown deal with a situation of only two transformers where the reflections at discontinuities may still be large, and so the theory explained here is least accurate for their case. The two methods are therefore, in a manner, complementary.

LEO YOUNG.

Electronics Division,
Ship & Shore Elec. Equip. Design,
Westinghouse Electric Corp.,
Baltimore, U.S.A.
1st March 1954.

[1] "A Wide-Band Waveguide Phase-Shifter", by G. J. Halford, *Proc. Instn elect. Engrs*, Part III, May 1953, Pp. 117-124.

Spectrum Equalization

SIR,—In his paper "Spectrum Equalization" (*Wireless Engineer*, May 1953, p. 112), Gouriet says that time delay makes it impossible to achieve ideal equalization, even within a restricted bandwidth. This is quite true, for the ideal equalization characteristic would have a transfer function which was the reciprocal of that of the network to be equalized and so would have a negative time delay.

Later, however, Gouriet says that "This article describes a means of producing the desired equalizing characteristic by adding to the response function its successive derivatives or integrals with respect to time with appropriate adjustment of the amplitudes and signs."

I would like to point out that, in fact, precisely the same restrictions apply to this method as to the conventional equalizer. Taking Fig. 8 of the paper, and including the inevitable capacitance across the transformer secondary and across the valve output, the exact transfer function has four poles and it is only at small values of p that it approximates to pMg_m, which is the Laplace transform of a differentiating network.

In view of this, Gouriet's claim (2) on p. 122 applies equally to any appropriate network and his claim (3) is erroneous.

In saying this, I do not wish to detract in any way from the importance and practical utility of Gouriet's work and I would like to stress much more fully than he did that his claims (1) and (4) have turned out to be very effective indeed in improving the reproduction of television systems. The flexibility obtained with it is of particular importance.

J. PETERS.

Nordwestdeutscher Rundfunk,
Hamburg.
8th March 1954.

SIR,—There was no intention to suggest in my article that the derivative method of equalization would compensate for 'real' time delay. This question is dealt with in Sec. 2.3.

It is, of course, true that the restrictions which apply to

linear circuit elements will apply also to any equalizer which incorporates such elements. Over a limited frequency band, however, derivatives may be extracted with any desired accuracy, and I would submit that if the error is too small to observe then one is entitled to regard such error as being non-existent over the frequency band concerned. Furthermore, it is not essential to use conventional circuit elements as the means for obtaining derivatives. For example, the differential of a signal may be obtained by subtracting from the signal a delayed version of it, in which case the differential becomes exact as the delay approaches zero. The transfer characteristic of such an arrangement has more zeros than poles and is therefore not subject to the same restrictions as a differentiating 'circuit', such as a transformer, or an RC network.

The fact that even-order derivatives will provide compensation for an attenuation characteristic without causing dispersion or introducing time delay, does, I consider, support claim (3). It might be argued that in practice, pure derivatives cannot be produced, but this is also true of pure resistance and pure reactance!

G. G. Gouriet.

B.B.C. Research Department,
Kingswood Warren, Surrey.
15th March 1954.

NEW BOOKS

Low Frequency Amplification

By Dr. N. A. J. Voorhoeve. 495 pp. + xv. Philips Technical Library, Cleaver-Hume Press, Ltd., Wright's Lane, Kensington, London, W.8. Price 50s.

Sound amplification has of recent years increased in importance; public address systems have grown in number and scope, and the appreciation of the nicer points of sound reproduction by engineers, musicians and the general public has increased, culminating in a demand for the highest fidelity. There is, however, a notable lack of books dealing with the subject in a thorough fashion. This book attempts to supply this need and the author sets himself a wide interpretation of his subject. About two-thirds of the book is concerned with audio-frequency circuitry, the remainder dealing with signal sources, loudspeakers and problems associated with complete amplification systems.

The author has had as his aim, a guide and reference book for practising sound engineers and for others with a less specialized technical training.

The book is one of the Philips Technical Library series and is a translation from the Dutch. It starts with a short discussion of principles followed by a chapter on valves, containing, by way of illustration, detailed descriptions of a number of Philips valves. Voltage amplifier circuits are next described followed by a chapter on power amplifiers; this last section, as might be expected, being particularly detailed. Chapters are devoted to feedback, components, and power supplies.

The last part of the book is concerned with the operation of complete amplifiers and their ancillary equipment. There is a useful chapter on principles of acoustics, including reference to the properties of the ear, and the effects of the characteristics of amplifier circuits on these properties. Input sources are described briefly, and loudspeakers a little more fully, while there are several chapters on aspects of complete sound reproduction systems, including a detailed chapter on amplification problems in buildings and other public places. The book ends with a description of measuring techniques and instruments, and a section of tables of symbols and terms. There is also a list of the most important formulae developed and used in the book, with references to the appropriate part of the text.

Books of this type are inclined to be somewhat uneven in treatment, having sections of great detail interspersed with discursive passages, and this book does not altogether escape this criticism, but has the compensating advantage of extensive bibliographies with each chapter. References in the text are numerous and the reader can easily amplify his reading about any subject of interest. Some of the information is summarized in the form of tables and graphs in the text and this, together with the general arrangement of the text, assists in easy reference to a particular subject. The style is, in general, simple and interesting, except for an occasional ambiguous or confusing passage. Throughout the book the author has illustrated his matter with numerous references to specific Philips products and, although he states in his preface that products of the leading manufacturers are equivalent, it may well be felt that unnecessary stress has been laid on Philips manufactures in some cases.

Unfortunately there is an unusually large number of small misprints and other errors and, although these are often not important in themselves, they give an impression of carelessness leading to a lack of confidence in the accuracy of the whole of the book. A typical example occurs on page 18 in the chapter on principles, where the treatment of the simple, but often misunderstood, question of the change in voltage gain of an amplifier with differing input and output impedances, is obscured by the substitution in two places of V_t^1 for V_t^2. It is difficult to understand how some of the errors have escaped the notice of the proof readers, for example on pages 396 and 397 cosh θ, sinh θ and tanh γl are printed as cos $h\theta$, sin $h\theta$ and tan $h\gamma l$ in a formula which is simply quoted, making the error difficult to spot by one unfamiliar with it. Even the symbols appendix is not free from such errors.

Some of the symbols in the text have unnecessarily complicated subscripts which do not help either the reader or the compositor; no less than four versions of the "approximately equal to" sign are used at different times, tending to suggest different orders of approximation in each case.

However, if these disadvantages are ignored, the book offers the practising sound engineer and the specialist in allied fields a useful reference book and refresher course in the wider aspects of sound amplification and reproduction.

F. L. W.

Einschwingvorgänge, Gegenkopplung, Stabilität. Theoretische Grundlagen und Anwendungen.

By Johannes Peters. Pp. 181 + xv, with 130 illustrations. Springer-Verlag, Reichpietschufer 20, Berlin, W.35. Price D.M.27.

The author is head of the research department of the Nordwestdeutscher Rundfunk in Hamburg, and also gives lectures at Hamburg University. The book is of a very mathematical character and might very well have been entitled "the mathematics of back-coupling." It originated in an attempt to develop the theory of feedback amplifiers and the conditions for stability by a strictly accurate method without the usual approximations. The contents are divided into five chapters, of which the first deals with the static and dynamic properties of linear systems; the reader soon finds himself in a world of complex frequencies, Cauchy-Riemann equations and Fourier and Laplace transforms. The second chapter is entitled "Transmission Factors of

Passive and Active Systems"; it deals very thoroughly with the analysis of networks and with the various types of matrices. The third chapter is entitled "Stability, Stability Criteria and Stabilizing"; the author says that it is difficult to determine on what factors the stability depends and he feels justified in devoting a separate chapter to it. The following statement in italics is rather amusing: "*Nyquist's stability criterium holds whenever it is possible to check it by a practical measurement.*" Considerable space is devoted to Nyquist's criterion. Here and in other places the author acknowledges his indebtedness to H. W. Bode's "Network Analysis and Feedback Amplifier Design."

The fourth chapter is devoted to the actual problem of the Feedback Amplifier; all the various types of feedback circuits are very fully discussed, especially from the point of view of their practical utility. The final chapter discusses the analogies between the mechanical and electrical systems and then the use of electro-mechanical converters of various types, including magnetostriction and piezo-electric. Although servo-mechanisms are not mentioned by name, they are closely associated with several items in this chapter. In addition to the index there is a very extensive list of books and papers on the subject. With the book there is issued a sheet of corrections.

Occasionally an example—usually numerical—is given to illustrate the mathematical treatment. The first one, which occurs on p. 6, seems to make a muddle of a simple problem. We give it in the original German. "Beispiel. Das Übertragungssystem sei ein Spannungsteiler, bestehend aus einer Reihenschaltung eines Widerstandes R, einer Induktivität L und einer Kapazität C mit der Spannungsquelle am Eingang. Die Ausgangsspannung wird im Leerlauf parallel zu dem aus Widerstand und Kapazität bestehenden Zweig abgenommen. Dann ist der Übertragungsfaktor $z(\omega) = (R + i\omega L)/(R + i\omega L + 1/i\omega C)$." That is to say, a potential divider consists of R, L and C connected in series across the input voltage, and the output voltage on open circuit is taken "parallel to the branch consisting of R and C"; i.e., presumably across R and C. The numerator of the output/input voltage will then be $R + 1/i\omega C$ and not $R + i\omega L$ as the author states, but the wording is strangely vague.

After several pages devoted to complex frequencies one finds a footnote on p. 13 explaining that frequency in the technical sense means the number of periods per second, but this is due to the fact that, as he explains on p. 2, he normally uses the word frequency to denote the angular frequency ω.

The book is excellently produced and will undoubtedly prove of great interest to anyone desirous of exploring the mathematics of feedback systems. G. W. O. H.

Radio Receiver Design

Part 1. Radio Frequency Amplification and Detection. By K. R. Sturley, Ph.D., B.Sc., M.I.E.E., Sen.M.I.R.E. 2nd Edition. Pp. 667 + xx. Chapman & Hall, Ltd., 37 Essex Street, London, W.C.2. Price 56s.

Anyone familiar with the first edition of this book is immediately struck by the relative bulkiness of the second. It actually has 667 pages as compared with 435. The eight chapters have the same headings as before and in broad outline cover much the same ground.

Chapter 1, on general considerations, has been rewritten and now occupies 49 pages instead of 16. The changes are mainly additional material on frequency modulation, pulse modulation and noise. The second chapter, on valves, is not much altered save for the addition of 25 pages on noise. Chapter 3, on aerials and aerial-coupling circuits, has also been expanded somewhat. Some information on folded-unipole and folded-dipole aerials has been added, as well as a section on slot aerials. The use of directors and/or reflectors is not mentioned, however, but a couple of pages on balance-to-unbalance transformers has been inserted and a lengthy section covering wave-traps is new.

Chapter 4, on r.f. amplification, is basically unchanged but has had some extra material inserted. There is more about r.f. resistance, Q, and self-capacitance of air-core coils and about mutual inductance, and there is much more about amplification at high radio frequencies.

Chapter 5, "Frequency Changing", has had the section on diode mixers greatly expanded, some nine pages extra on noise have been added and nearly three chapters on the Synchrodyne, but otherwise it is little changed. Chapter 6, "Oscillators", is also not greatly changed, except for the (lengthy) section on tracking.

Chapter 7, on intermediate-frequency amplification, has been considerably expanded and, among other things, there are over 40 extra pages on crystal filters. The last chapter, on detectors, has been relatively little altered. There are several new appendixes.

The book, as a whole, is a very valuable reference book for the designer and the new edition is a considerable improvement on the first. It is very mathematical but the mathematics is all of a quite simple nature and involves little more than ordinary algebra and the j notation, which is explained in an appendix.

A somewhat surprising omission, in view of the way in which air-core coils are treated, is the lack of corresponding data on dust-iron core types, for they are now, by far, the more widely used. They do not appear in the index at all and references to them in the text are confined to a few isolated sentences. While it would undoubtedly have been helpful to have had a discussion on the characteristics of dust-iron core coils in the book, the omission is really rather a minor one in view of its general excellence.

The revision has been very thoroughly carried out and what has always been a most useful and reliable book has been made even more valuable. W. T. C.

Complex Variable Theory and Transform Calculus (2nd Edition)

By N. W. McLachlan, D.Sc.(Eng.). Pp. 388. Cambridge University Press, 200 Euston Road, London, N.W.1. Price 55s.

The first edition was published in 1939 under a slightly different title; the book has been completely revised. Some pure mathematicians and physicists who reviewed the first edition of this book apparently regarded the degree of rigour as inadequate. They must in spirit be the descendants of the Cambridge mathematicians who obstructed Heaviside. It is the attitude of Heaviside that is needed by the technologist who uses mathematics to solve technical and industrial problems to-day. In such problems, most of the difficulty consists in knowing what kind of solution to seek, expect or guess. Operational calculus increases one's power of guessing effectively by many decibels. The process of verifying whether a guess is correct or not is relatively easy. In most cases it will show either that the guess was correct or that it was radically wrong. In the latter case there may be some indication of how to make a better guess. If the reason for error in a solution suggested by operation calculus cannot be easily found, the pure mathematician should be consulted. The technologist only needs to be warned of important pitfalls.

The first seven chapters deal with functions of a complex variable, Cauchy's theorem, singularities, the calculation of residues, the Bromwich contour, branch points, and differentiation and integration under the integral sign; Chapter 5 devotes special attention to

gamma, error and Bessel functions. The essential groundwork of complex-variable theory is thus covered, with emphasis on the parts relevant to operational calculus, and particularly on the determination of residues. Chapters 8-10 discuss transform calculus, the Mellin inversion theorem, solution of ordinary linear differential equations, discontinuous functions, impulses and frequency spectra. The remaining chapters deal mainly with applications, but the solution of partial differential equations in two variables by transformation into ordinary differential equations involving the operator p is discussed in Chapter 13; 102 examples for the reader are included, with answers. There is also a bibliography of 273 references. Proofs of many results considered too long or difficult for complete inclusion in the text are indicated in outline with perhaps rather too many references to specific pages of books or papers in this bibliography.

The 'p-multiplied Laplace transform' is used consistently, so that transforms obtained agree with those found by Heaviside's method; for example, $t^n/n!$ corresponds to $1/p^n$, and unit step to unity; all functions of time involved are assumed zero for negative times. If $h(t)$ corresponds to $f(p)$, then $f(j\omega)/j\omega$ is the Fourier transform of $h(t)$, and $|f(\omega)/j\omega|$ is its spectrum.

In electrical applications, the general procedure required is: Let $\phi(p)$ be the transform of the input $h(t)$. The effect of the circuit is to provide an impedance $Z(p)$ so that the output is the transform of $\phi(p)\ Z(p)$ in the time world. The contribution of a resistance R to impedance is R, that of an inductance L is Lp, and that of a capacitance C is $1/Cp$; these can be combined for series and parallel circuits by well-known rules. Now frequently $\phi(p)\ Z(p)$ contains a rational function of p as a factor. In practice, this will usually have only simple poles, and its resolution into partial fractions is then straightforward, and will often reduce $\phi(p)\ Z(p)$ to a combination of functions whose transforms are already known. This reduction to partial fractions is not adequately emphasized in the present book. The deficiency is mitigated to some extent by another book by Professor McLachlan.[1] In collaboration with Professor Humbert,[2] he has also provided a first-class list of transforms with brief indications of restrictions on their validity.

The reduction of $\phi(p)\ Z(p)$ to a combination of known forms, if necessary with the aid of partial fractions or of general rules such as the 'convolution integral', is surely preferable to the use of the Mellin inversion theorem in most cases of interest to the technologist, who is likely to find the application of this theorem difficult. He needs encouragement to investigate 'p' as a high-powered solution detector. Even mistakes made in the early stages will mostly be of the profitable kind by means of which he learns.

The typography is of the high standard one expects from the Cambridge University Press, but one of the few misprints, on page 195, is unfortunate. In equation (3), L should be replaced by R; (3) would then agree with the well-known formula (6) on the same page. A reader eager to apply what he has learnt in Chapters 1-10 might find his confidence sapped for some time by this inconsistency.

The book contains a great deal of useful information, particularly on the determination of residues and in the bibliography, and should be freely available in libraries, but it will be of more value as a book of reference to those familiar with operational calculus than as a textbook to those initially ignorant of it. J. W. H.

[1] N. W. McLachlan, "Modern Operational Calculus with Applications in Technical Mathematics", Macmillan, London, 1948.
[2] N. W. McLachlan, Bibliography, p.376, ref. 235a, and P. Humbert, "Formulaire pour le Calcul Symbolique". Mém. des Sci. Math. Fascicule 100 (1941), 2nd edn (1950).

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(Communication from the National Physical Laboratory)

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Date 1954 Feb.	Frequency deviation from nominal: parts in 10^8		Lead of MSF impulses on GBR 1000 G.M.T. time signal in milliseconds
	MSF 60 kc/s 1429-1530 G.M.T.	Droitwich 200 kc/s 1030 G.M.T.	
1	−1·2	NM	+36·8
2	−1·3	NM	+34·8
3	−1·3	−2	+33·8
4	NT	−1	NT
5	−1·2	−1	+32·2
6	NM	−2	NM
7	NM	0	NM
8	−1·3	0	+28·2
9	−1·3	0	+27·3
10	−1·3	+2	+26·2
11	−1·3	+1	+23·7
12	−1·2	+1	+22·1
13	−1·2	0	NM
14	−1·2	+1	NM
15	−1·1	+1	+16·7
16	−1·1	+1	+15·4
17	−1·1	0	+13·4
18	−1·1	−2	+11·5
19	−1·2	−1	+ 9·8
20	−1·2	−2	NM
21	−1·2	−2	NM
22	−1·2	−3	+ 6·8
23	−1·2	−1	+ 5·6
24	−1·3	−3	+ 5·2
25	−1·2	−1	+ 4·8
26	−1·1	−1	+ 4·3
27	−1·1	−1	NM
28	−1·2	−1	NM

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QQV03-20	R.F. Power Double Tetrode	B7A	12.6 0.65	6.3V 1.3A	600	2×10	250	2×1.5	2×55
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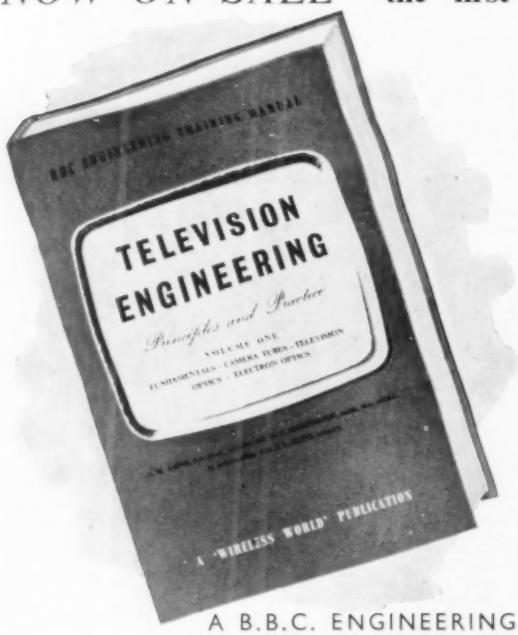

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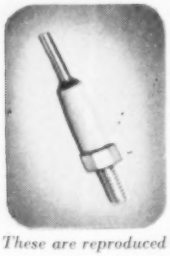

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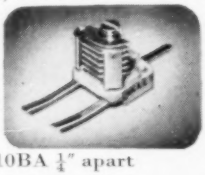

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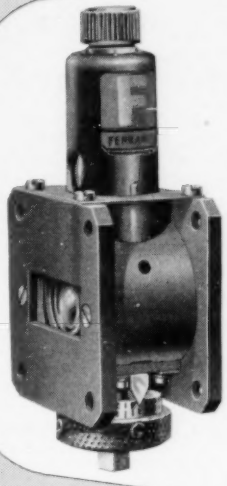

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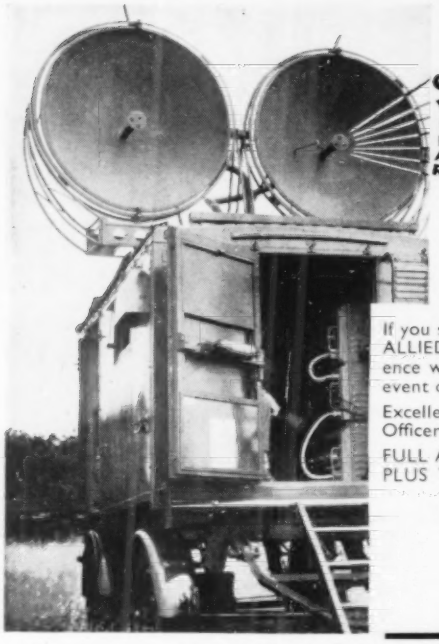

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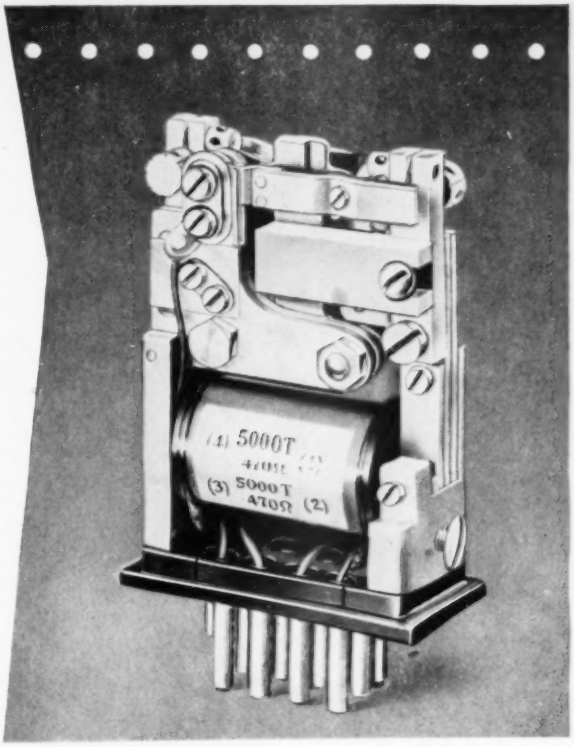

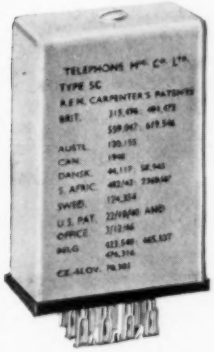

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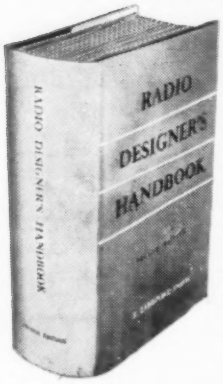

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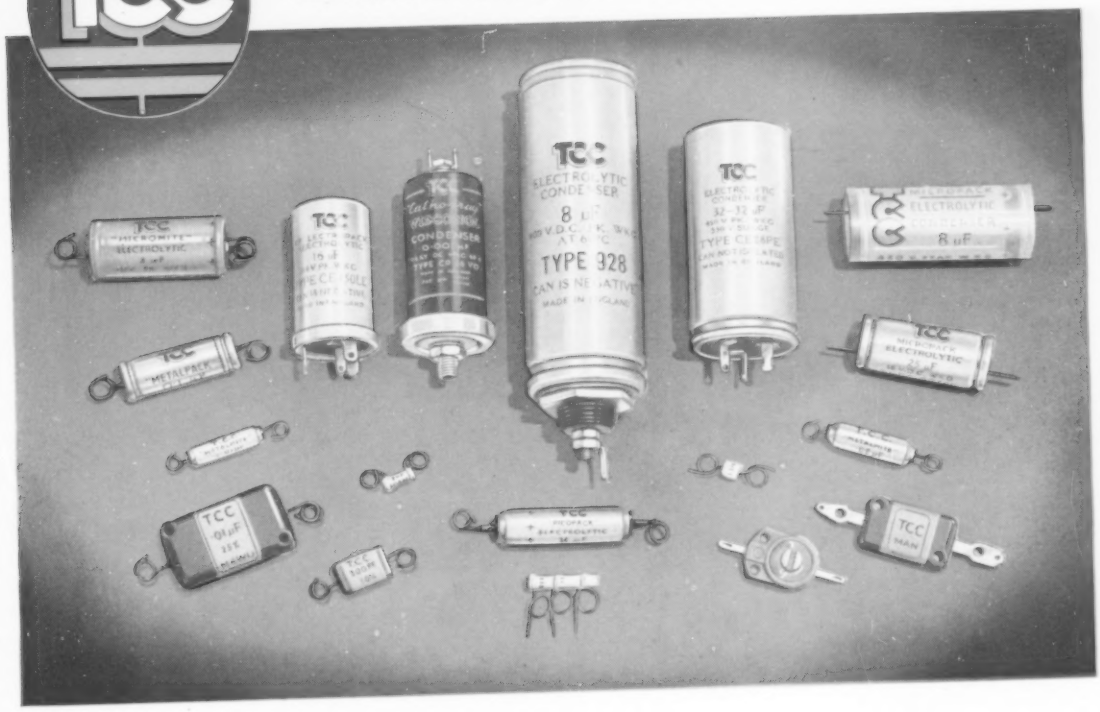

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